Addis Ababa University

Addis Ababa Institute of Technology

School of Electrical and Computer Engineering

Performance Analysis and Evaluation of Pilot-based Channel Estimation Techniques for Broadcasting Systems

By: Hailemariam Teklay

Thesis Submitted To Addis Ababa Institute of Technology in Partial Fulfillment of the Requirements for the Degree of Master of Science in Electrical and Computer Engineering (Communication Engineering)

Advisor: Dr. -Ing. Dereje Hailemariam

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Addis Ababa, Ethiopia

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Approval by Board of Examiners

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DECLARATION

I, the undersigned, declared that this MSc thesis is my original work, has not been presented for fulfillment of a degree in this or any other University and all sources and materials used for the thesis are duly acknowledged.

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This thesis work has been submitted for examination with my approval as a University Advisor.

Dr. -Ing. Dereje Hailemariam ________________________

Advisor's Name Signature

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Abstract

Terrestrial broadcasting channel is unknown and prone to time dispersion, which causes inter symbol interference (ISI) and fading. This unknown channel must be estimated based on analysis of received signal and training pilots before equalization at the receiver side.

In this thesis work the performance of two pilot-based channel estimation algorithms; namely, least square (LS) and Minimum Mean Square Error (MMSE) with corresponding channel interpolation techniques are investigated for broadcasting system applications. This analysis is part of a joint project between Information Networks Security Agency (INSA) of Ethiopia and the Addis Ababa University (AAU) that intends to implement advanced digital receiver for a certain application. Mean square error (MSE), Bit error rate (BER), rate of convergence, stability and complexity are used as performance metrics.

The MATLAB simulation result shows the performance of MMSE estimator is much better than LS estimator despite its complexity in both 4-QAM and QPSK modulation. To estimate the channel coefficients Typical Urban channel - 6 paths (TU6) channel model is used.

From all the simulation results, LS and MMSE algorithms are able to extract channel coefficients but MMSE has better performance. Therefore, the thesis recommends MMSE algorithm for channel estimation implementation by INSA.

Key words: - *Channel estimation, Channel interpolation, Pilots, DVB-T2, LS and MMSE*

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List of Abbreviations and Notations

3GPP	3rd Generation Partnership Project
3GPP2	3rd Generation Partnership Project2
ADR	Astra Digital Radio
AWGN	Additive White Gaussian Noise
BBHEADER	Baseband Header
BC	Broadcasting
BCH	Bose-Chaudhuri-Hocquenghem
BER	Bit Error Rate
BICM	Bit Interleaved Coded Modulation
CD3	Coded Decision Directed Demodulation
CE	Channel estimation
CIR	Channel Impulse Response
CMA	Constant Modulus Algorithm
COFDM	Coded Orthogonal Frequency Division Multiplexing
CR	Coding Rate
CRC	Communications Research Center
DDA	Decision-Directed Algorithm
DFT	Discrete Fourier Transformation
DigiTAG	Digital Terrestrial Television Action Group
DSP	Digital Signal Processing
DSR	Digital Satellite Radio
DTT	Digital Terrestrial Television
DVB-C	Digital Video Broadcasting – Cable
DVB-S	Digital Video Broadcasting – satellite
DVB-SH	Digital Video Broadcasting - Satellite services to Handhelds
DVB-T	First Generation Digital Terrestrial Television
DVB-T2	Second Generation Digital Terrestrial Television
EM	Expectation-Maximization
ETSI	European Telecommunication Standards Institute

FDM	Frequency-Division Multiplexing
FEC	Forward Error Correction
FFT	Fast Fourier Transformation
GA	Generic Algorithm
GI	Guard Interval
GIF	Guard-Interval Fraction
GS	Generic Streams
GSM	Global System for Mobile Communications
HDTV	High Definition Television
HEM	High Efficiency Mode
HOS	Higher-Order Statistics
IFFT	Inverse Fast Fourier Transform
ISI	Inter symbol Interference
LAN	Local Area Network
LDPC	Low Density Parity Check
LMS	Least Mean Squares
LOS	Line of sight
LS	Least Square
MFN	Multi Frequency Network
MHP	Multimedia Home Platform
MI	Modulator Interface
MIMO	Multiple Input Multiple Output
MISO	Multiple Input Single Output
MLSE	Maximum-Likelihood Sequence Estimator
MMSE	Mean Minimum Square Error
MMSE	Minimum Mean Square Error
MPEG	Motion Picture Experts Group
MSE	Mean Square Error
NLMS	Normalized Least Mean Squares
NLOS	Non Line of sight
NM	Normal Mode

OFDM	Orthogonal Frequency Division Multiplexing
PAPR	Peak to average Power Ratio Reduction
PDP	Power delay profile
PI	Pedestrian Indoor
PLP	Physical Layer Pipe
PO	Pedestrian Outdoor
QAM	Quadrature Amplitude Modulation
RA	Typical Area
RF	Radio-Frequency
RLS	Recursive Least Square
RMS	Root Mean Square
RS	Reed-Solomon
SCR	Spectral Coherence Restoral
SDA	Steepest Descent Algorithm
SFN	Single Frequency Network
SISO	Single Input Single Output
SNR	Signal to Noise Ratio
SOS	Second Order Statistics
STBC	Space-Time Block Codes
STTC	Space-Time Trellis Codes
TBCE	Training Based Channel Estimation
TDT	Terrestrial Digital TV
TFS	Time Frequency Slicing
TI	Texas Instruments
TS	Transport Stream
TU	Typical Urban
TV	Television
UHF	Ultra-High Frequency
UP	User Packet
VBR	Variable Bit Rate
VHF	Very High Frequency

Wi-MAX	Worldwide Interoperability for Microwave Access

List of Notations

$(.)^{H}$	Matrix Hermitian
$(.)^{-1}$	Matrix inverse
$(.)^{T}$	Matrix (vector) transpose
$(.)^{*}$	Complex conjugate
$E\{.\}$	Mathematical expectation
I_m	$m \times m$ Identity matrix
Rx	Receiver
Tx	Transmitter
μ	Step size

Chapter One

Introduction

1.1 Background

The wireless evolution has been stimulated by an explosive growing demand for a wide variety of high quality of services in voice, video, and data. This rigorous demand has made an impact on current and future wireless applications, such as digital audio/video broadcasting, wireless LANs, Wi-MAX, cognitive radio, 3GPP and 3GPP2 Long Term Evolution, and wireless sensor networks [1]. These advanced applications in which the transmitted signal disperses over the time and the frequency domains, show the need for highly-developed signal processing algorithms. In particular, one of the main challenges in the wireless communication is a wireless channel that suffers from numerous physical impairments due to multipath propagation, interference from other users or layers, and the time selectivity of a channel [1].

There are number of techniques used to mitigate these challenges such as equalization, diversity and channel coding. These techniques can be used independently or in tandem to improve received signal quality. In time varying multi-path fading channel that exists in mobile communications environment lead to severe ISI. Channel equalization is one of the mitigation techniques used to compensates for ISI created by multipath within time dispersive channels [2].

Before equalization channel should be estimated based on analysis of received signal and the estimated channel coefficients are given as input for channel equalization for further received signal compensation as it shown in Figure 1.1 below. Channel estimator estimates channel fading coefficients by periodically adjusting an adaptive linear filter according to an algorithm so as to minimize the error between the output of the channel estimator and the received signal.

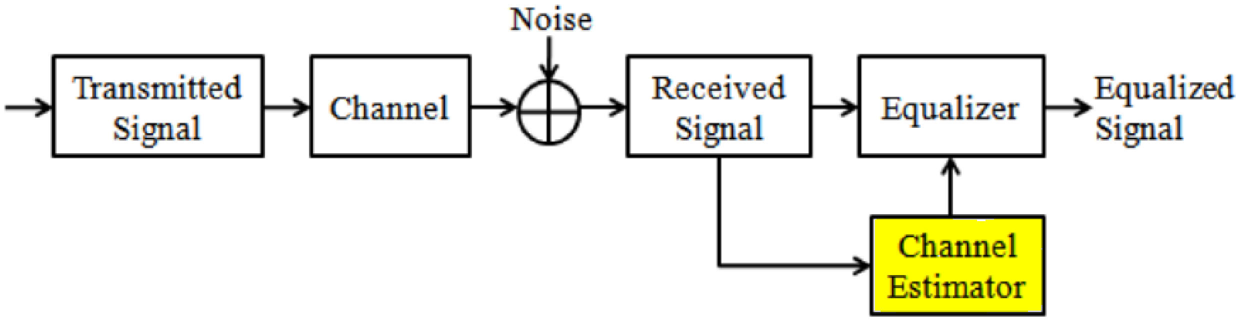


Figure 1.1 Block diagram of channel estimation and channel equalization *[3]*

Several channel estimation techniques have been proposed to improve the estimation process and reduce the computational complexity by exploiting certain characteristics of the channel model. The most common techniques are pilot-based /non-blind, semi blind and blind channel estimation techniques.

Non-blind (pilot-based) channel estimation techniques use block – type or comb-type pilots as a reference signal for training. This techniques is the most common and simple technique but bandwidth efficiency decreases with the size of pilots. Blind techniques, that do not require any training symbols, achieve high system throughput with high computational complexity. Semi-blind schemes require less computational complexity than blind methods and fewer training symbols than training-based methods, making them attractive for practical implementation.

This thesis work mainly focus on pilot-based channel estimation techniques since this technique is applicable for Raleigh fading channels and in terrestrial environment practical implementation.

1.2 Overview on Digital Broadcasting Systems

Broadcasting is the distribution of audio and/or video content to a dispersed audience via any electronic mass communications medium, but typically one using the electromagnetic spectrum (radio waves), in one to many model [4].There are many digital broadcasting standards the most familiar in Europe and other countries is Digital video Broadcasting (DVB).

DVB standards family is a set of internationally accepted open standards for digital television. DVB standards are developed by the DVB Project, an international industry consortium with more than 270 members. Its objective is to agree specifications for digital media delivery systems, including broadcasting. It is originally of European origin but now worldwide [6].The standard provide for delivery of the data by terrestrial (DVB-T), cable (DVB-C) and satellite (DVB-S) infrastructures [3].

DVB-S is Digital broadcasting systems for television, sound and data services; framing structure, channel coding and modulation for 11/12 GHz satellite services. The modem standard for satellite broadcasts with various data rates, band width requirements, and error correction capabilities [4]. Used for long distance~36000km downlink but channel characteristics is unknown because of weather condition (rain, clouds).

DVB-C is Digital broadcasting systems for television, sound and data services; framing structure, channel coding and modulation for cable systems. The modem standard for cable broadcasts with various data rates, levels of noise immunity, and band width requirements [4]. Used for short distance communication but channel characteristics constant no need of channel estimation and equalization.

DVB-T Framing structure, channel coding and modulation for digital terrestrial television. Terrestrial transmission path is subject to numerous impacts such as echoes and multipath reception, AWGN and Doppler shift in case of mobile reception that needs channel estimation and channel equalization at the receiver side [5]. Throughout this work analysis and implementation of channel estimation techniques is proposed on moderate and very slow mobility DVB receivers, focusing on the DVB-T2.

1.3 Statement of the Problem

The wireless channel in mobile radio poses a great challenge as a medium for reliable high speed communications. When a radio signal is transmitted through a wireless channel it suffers various types of distortions. Hence, the receiver obtains a linear superposition of the signals transmitted by all the users, attenuated by arbitrary factors and delayed by an arbitrary amount. The effect of the channel on the transmitted signal should be estimated correctly before equalization is made. Many scholars and researchers invest their time to solve the problem and published different journals and articles. But locally in our country there is demand to make analysis of such mitigation techniques for further implementation in a real time. Information Networks Security Agency (INSA) of Ethiopia is working with the Addis Ababa University (AAU) on an industrial project that aims to implement efficient receiver to mitigate the effects of multipath channel on the transmitted signal for a certain application. Hence, the motivation of the thesis is identifying an optimal pilot-based channel estimation method for broadcasting system and thus, address the need of the industry.

1.4 Objective of the Thesis

The thesis is aimed to achieve the following general and specific objectives.

1.4.1 General Objective

The general objective of the thesis is performance analysis and evaluation of pilot-based channel estimation techniques using MMSE and LS algorithms for broadcasting systems.

1.4.2 Specific Objectives

The specific objectives of the thesis are:

- To study different channel estimation techniques and select appropriate estimation technique.
- To compare the performance between LS and MMSE channel estimation algorithms.
- To make performance analysis of different channel interpolation techniques.
- To develop performance and complexity analysis of different channel estimation algorithms.
- To recommend/propose the best estimation techniques for real practical implementation.

1.5 Contribution of the Thesis

In wireless communication systems the source data is encoded, modulated and transmitted over the unguided channel (air). During transmission, the signal encounters lot of impairments such as noise, fading and interference. At the receiver end, the effect of the channel should be first estimated and then compensated (equalized) so as to regain the transmitted signal. In order to meet these requirements, the receiver must employ better performance channel estimation algorithms and techniques. This thesis work make performance analysis on channel estimation algorithms for broadcasting systems and the end results and analysis of this thesis work will be used as an input to a practical implementation of reliable and efficient receiver for the industry.

1.6 Scope and Limitation

The scope of the thesis covers study and performance analysis of LMS, RLS, MMSE and LS pilot based estimation algorithms but the simulation considers only MMSE and LS estimators. The performance analyses and recommendation are based on BER, rate of convergence, stability and complexity using 4-QAM and QPSK modulation. The simulation is done using entirely the MATLAB simulation tool considering only random data and image file as inputs.

This thesis work will not include the performance analysis of blind and semi-blind types of channel estimation techniques.

1.7 Methodology

The methods used to achieve the desired objectives of this thesis are as follows.

1. **Literature review**:

   Include reading books, journals, articles, simulation tools, and other resources related to channel estimation techniques (i.e., blind, non-blind and semi-blind) and different channel estimation algorithms such as LS, LMS, RLS, MMSE, CMA and LMMSE

2. **System modeling**:

   Involves modeling of the overall system in using the appropriate channel models and estimation algorithms for broadcasting systems.

3. **Simulation**:

   Simulating the modeled system using MATLAB, performance comparison of different channel estimation algorithms and then incorporating implementations of selected channel estimation algorithms.

4. **Performance comparison**:

   Include comparing the BER-SNR, MSE-SNR performance, computational complexity and stability of the CE algorithms and channel interpolation techniques.

5. **Analysis and interpretation of the results**:

   Finally, the results are interpreted and conclusion is drawn based on the results obtained.

1.8 Literature Review

To minimize the effect of multipath channel on the transmitted signal and mitigate ISI, a lot of researches have been done which most of them have mainly focused on joint channel estimation and equalization techniques. Some of the reviewed related works are given as follows. **Marco Rotoloni**, proposes channel estimation techniques for OFDM with application to Digital Video Broadcasting Standards. The paper deals with timing and frequency synchronization, channel estimation and data detection. The channel estimation was focus on channel estimation using known pilot symbols. Transmission of pilot symbols on a sub-set of sub carriers in an OFDM system allows for an efficient channel estimation at the receiver by means of the LS method. Usually, for these systems channel estimation is performed on pilots and then interpolated over the time and the frequency axis. Performance results was done with reference to the DVB-T and DVB -T2 standards [4].

N. V. Ratnam, proposes DSP implementation of channel estimation algorithms for OFDM systems. This paper focuses on channel estimation with different interpolation approaches and adaptive algorithms OFDM system. The channel estimation based on the block - type, comb-type and lattice-type structure is studied. LS and MMSE channel estimation techniques were implemented using C6713 DSP of TI and LMS, NLMS and RLS were implemented and tested using Simulink [6].

Asmamaw Getu, Introduces genetic algorithm generic algorithm for joint channel estimation and data detection in application to multi-user MIMO. Several trained and blind channel estimation techniques and methods were studied. GA-based channel estimator and GA-based data detector are designed independently using selection, crossover, mutation, and replacement operators. Optimum parameter values are selected using observation of change in fitness values with change in parameter values. Performance comparisons between different CE and DD methods are made by using bit error rate BER and MSE values at specific SNR points [7].

S**ajjad Ahmed Ghauri, SherazAlam, M. Farhan Sohail, and AsadAli, FaizanSaleem**, cooperatively publishes international journal on Implementation of OFDM and Channel Estimation using LS and MMSE estimators. Their paper highlights the channel estimation technique based on pilot aided block type training symbols using LS and MMSE algorithm, This paper starts with comparisons of OFDM using BPSK and QPSK on different channels, followed by modeling the LS and MMSE estimators on MATLAB. Depending on results obtained conclude that LS algorithm gives less complexity but MMSE algorithm provides comparatively better results [8].

In review of the literatures it is solid clear that various research works has already been done on pilot-based channel estimation techniques. However, in my survey the papers are done on joint channel estimation and channel equalization do not clearly put channel estimation explicitly.

1.9 Thesis Organization

The work of this thesis is organized in to Six Chapters. Chapter one presents the introduction, methods and objectives of this thesis work. In Chapter two, we are going to see some theoretical parts of Wireless channels models and propagation models which are useful to model the thesis work and in Chapter three, channel estimation techniques: non-blind, blind and semi blind channel estimation techniques including corresponding channel estimation algorithms and channel interpolation techniques are presented. Then in the fourth Chapter we will see the broadcasting systems particularly DVB-T2 standards and basic parameters. Chapter five contains simulation setup and Simulation results. Finally, the last Chapter contains conclusions and recommendations for future works. At the end references are included.

Chapter Two

Channel Models for Terrestrial Broadcasting Systems

2.1 Multipath Propagation

In wireless telecommunications, multipath is the propagation phenomenon that results in radio signals reaching the receiving antenna by two or more paths. Causes of multipath include atmospheric ducting, ionospheric reflection and refraction, and reflection from water bodies and terrestrial objects such as mountains and buildings [1].

Multipath causes multipath interference including constructive and destructive interference, and phase shifting of the signal. Destructive interference causes fading. Where the magnitudes of the signals arriving by the various paths have a distribution known as the Rayleigh distribution, this is known as Rayleigh fading [9].

In broadcasting systems multipath causes jitter and ghosting, seen as a faded duplicate image to the right of the main image [10]. This happens when a signal arrives at the receiving antenna from multiple paths or direction. These multiple signals add up at the antenna, creating multiple images or "Ghosts" on the TV screen. This distortion is most pronounced in densely populated cities particularly those with high rise buildings [10].Therefore, for proper reception of terrestrial broadcast multi path the distortion, or Ghosts should be estimated and compensated using channel estimation and channel equalization techniques.

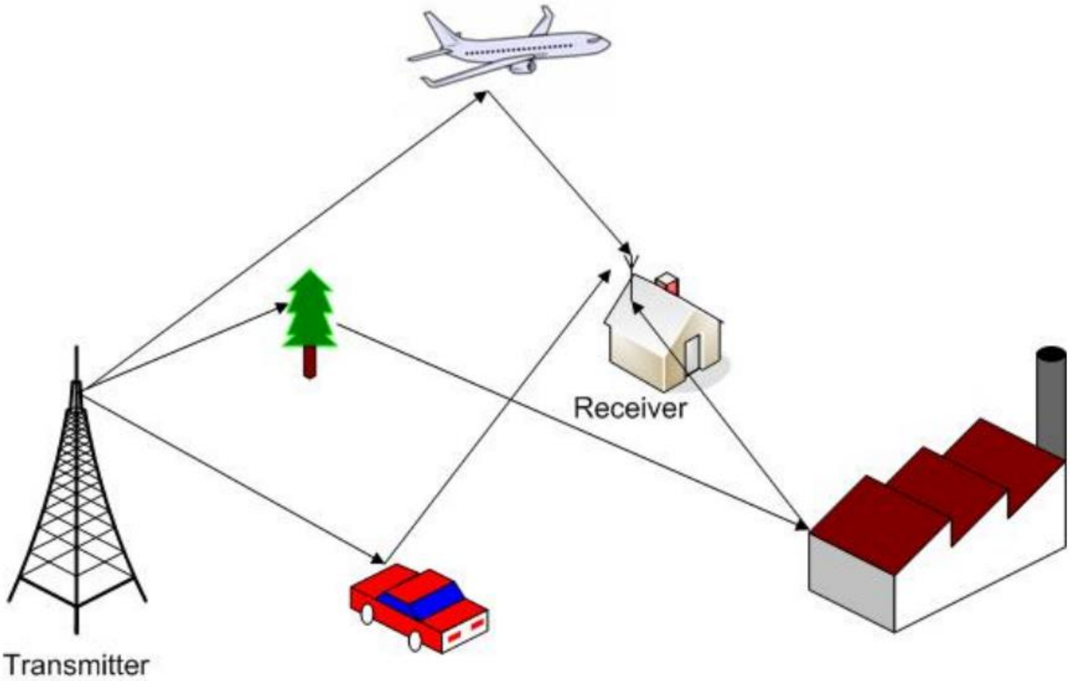


Figure 2.1 Multi- path transmission in a broadcasting application *[5]*

As it is shown in Figure 2.1 above there are many factors that impact signal propagation in the broadcasting system: reflection, diffraction, and scattering. The transmitted signal generally propagates to the receiver antenna through many different paths, thus, due to reflectors and scatterers as cars, buildings, trees, the receiver antenna will receive multiple delayed and scaled copies of the transmitted signal leading to temporary dispersion. On the other hand, the movement at the receiver produces frequency dispersion [9]. These effects lead to frequency and time-domain selectivity, respectively. In order to determine the effects that are present in the signal, it is necessary to consider their characteristics and the ones of the channel, such as the coherence time and bandwidth. In the following sections we will see different channel fading and some possible channel models for both large and small scale fading channels.

2.2 Fading Parameters

The multipath fading channel parameters are delay spread, coherence bandwidth, Doppler spread and coherence time. The first two are used to describe multipath mobile channel dispersion in time, and the latter two are used to describe multipath mobile channel dispersion in frequency [11].

2.2.1 Parameters of time Dispersion and Frequency Selectivity

Time dispersion and frequency selectivity are effects generated by superposition of the different multipath delay signals, depends on the geometric relationship between the transmitter, receiver and the surrounding physical environment. These two effects occur simultaneously, but in different forms [11]. Time dispersion is reflected in the time domain, and frequency selectivity reflected in the frequency domain. Time dispersion is a signal to the transmitter unfolded along the time axis, so that the duration of the received signal is longer than the signal sent. The frequency selectivity refers to the transmitted signal is filtered, the signal components of different frequencies in the range with different fading. It means when components are very close in frequency, and their decline is also very close, but when far apart in frequency, and their decline varies widely [11].

Delay spread: - In multipath propagation conditions, the received signal will produce delay spread. Parameters used to describe time extended are the average additional delay $\bar{\tau}$, RMS delay spread σ_τ and maximal delay spread. The first two parameters are related to power delay profile $p(\tau)$. PDP is a function of additional delay based on fixed time delay τ_0 by getting average of local instantaneous power delay [11].

Average additional delay is the first order matrix of PDP and written as

$$\bar{\tau} = \frac{\sum_{k=1}^{N} \alpha_k{}^2 \tau_k}{\sum_{k=1}^{N} \alpha_k^2} = \frac{\sum_{k=1}^{N} p(\tau_k) \tau_k}{\sum_{k=1}^{N} p(\tau_k)} \tag{2.1}$$

Where α_k is attenuation factor of k^{th} path, $p(\tau_k)$ is the relative power of multi-path fading at τ_k, respectively.

RMS delay spread is square root of second order matrix of PDP.

$$\sigma_\tau = \sqrt{E(\tau^2) - (\bar{\tau})^2} \tag{2.2}$$

$$E(\tau^2) = \frac{\sum_k^N \alpha_k{}^2 \tau_k{}^2}{\sum_k^N \alpha_k^2} = \frac{\sum_k^N p(\tau_k)\tau_k{}^2}{\sum_k^N p(\tau_k)} \tag{2.3}$$

Coherence bandwidth:- is used to characterize the channel in the frequency domain similar to the delay spread parameters in the time domain, Coherence bandwidth, B_c, is a statistical measure of the range of frequencies over which the channel can be considered "flat" meaning that a channel which passes all spectral components with approximately equal gain and linear phase. Equivalently, coherence bandwidth is the range of frequencies over which two frequency components have a strong potential for amplitude correlation [12].Coherence bandwidth is a determined value from RMS delay spread and could be written as below.

$$B_c = \frac{1}{\tau_{max}} \tag{2.4}$$

This is important since a signal having a larger bandwidth (B_s) than B_c is severely distorted for a 0.9 correlation $B_c = \frac{1}{50\sigma_\tau}$

2.2.2 Parameters of Frequency Dispersion and Time Selectivity

Due to the movement of the mobile station, the phenomenon of Doppler shift occurs, as frequency dispersion, and makes the channel time varying. Doppler spread and coherence time are two important parameters to describe the mobile channel frequency dispersion and time-varying characteristics, and there is the inverse relationship between them [11].

Doppler shift

Movement causes shift in signal frequency. When the stations are moving, the received signal frequency is different than the original signal frequency. The variation of the frequency of the received signal with time, caused by the relative motion between the transmitter and the receiver is called the Doppler shift.

Suppose a mobile is moving at the speed of V Doppler shift is given by

$$fd = \frac{V}{\lambda}\cos\theta = f_c\frac{v}{c}\ \cos\theta = f_m\cos\theta \tag{2.5}$$

Where v is the speed of mobile station, λ radio wavelength, f_c is carrier frequency of transmitter, c speed of light ,θ is the angle between radio and mobile station f_m is the maximum Doppler frequency shift.

From the equation we can see that if the movement is towards the signal generator, the Doppler shift is positive otherwise it is negative.

Doppler spread

A channel shows a time varying nature when there is a movement in either the source or destination or even objects in the middle. Doppler spread (B_D) is the measure of maximum broadening of the spectrum due to Doppler shift. Thereby B_D is $f_m = \frac{V}{\lambda}$ where f_m is the maximum Doppler shift. It is a frequency range, defined as the spectrum of the received signal spectrum is not equal to zero.

Coherence time

Coherence time T_c is the time domain dual of Doppler spread and is used to characterize the time varying nature of the frequency dispersive-ness of the channel in the time domain. The Doppler spread and coherence time are inversely proportional to one another [12]. That is

$T_c = {}^{1}/_{f_m}$

2.3 Fading Channel Types

According to the effect of multipath on the transmitted signal, fading is classified as large scale fading and small scale fading. Large-scale fading is manifested by the mean path loss and shadowing in which the received signal power varies gradually due to signal attenuation determined by the geometry of the path profile. On the other hand, small-scale fading refers to rapid variation of signal levels due to the constructive and destructive interference of multiple signal paths (multi-paths) when the mobile station moves short distances [13].

Depending on the relative extent of a multipath, frequency selectivity of a channel is characterized (e.g., by frequency-selective or frequency flat) for small-scaling fading. Meanwhile, depending on the time variation in a channel due to mobile speed (characterized by the Doppler spread), short-term fading can be classified as either fast fading or slow fading.

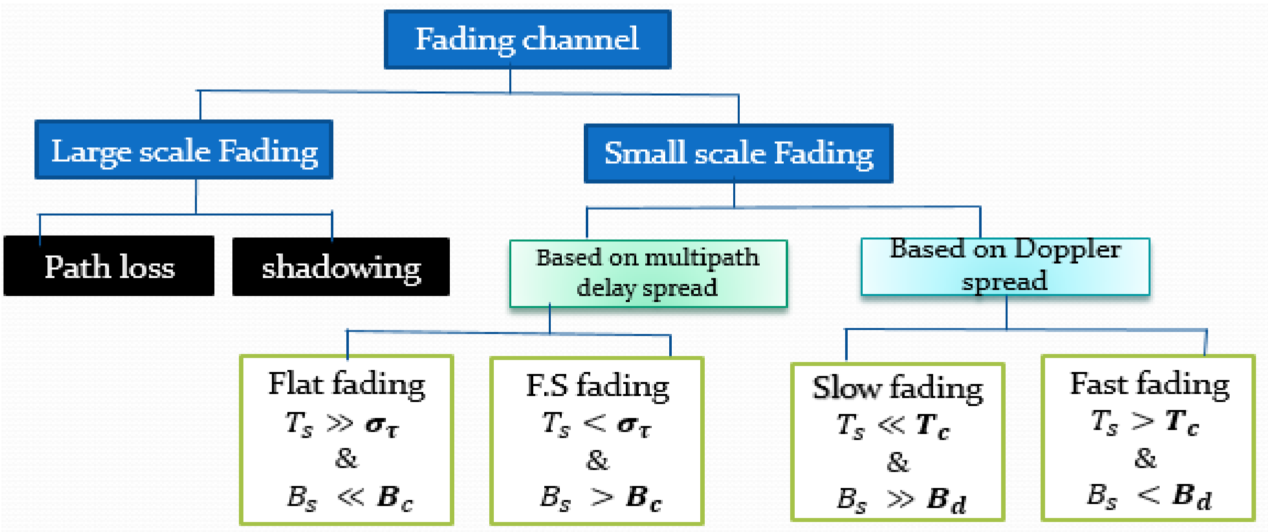


Figure 2.2 Classification of fading channels *[2, 10]*

2.3.1 Large-Scale Fading

Large-scale fading occurs as the mobile moves through a large distance and caused by path loss of signal as a function of distance and shadowing by large objects such as buildings, intervening terrains, and vegetation. Shadowing is a slow fading process characterized by variation of median path loss between the transmitter and receiver in fixed locations. In other words, large-scale fading is characterized by average path loss and shadowing [13].

Path loss is the reduction of an electromagnetic wave energy when propagates from a transmitter to a receiver over the air [14]. One of simple channel models is Friis transmission model given in Equation 2.6.

Shadowing is the loss of field strength typically contributed to a diffracted wave emanating from an obstacle between transmitter antenna and receiver antenna. As passing through a shadow area requires considerable time, the name 'slow fading' is commonly used. The shadow effect is modeled with a log-normal distribution of the mean signal [9].

Large-scale fading propagation models are used at the physical layer to predict the mean signal strength for an arbitrary transmitter-receiver separation distance. The free-space propagation model and the lognormal one are two generic propagation models that are often used as a basis for specific models [10].

2.3.1.1 Free-Space Propagation Model

This channel model is an ideal model used to compute the received signal strength when there is a direct LOS between a transmitter and a receiver unit, placed at distance d between them, without any obstacles near the line of sight. Free space model predicts that received power decays as a function of the Transmitter-Receiver separation distance raised to some power (i.e. a power law function).The free space power received by a receiver antenna which is separated from a radiating transmitter antenna by a distance d, is given by the Friis free space equation:

$$P_r(d) = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2 L} \tag{2.6}$$

Where P_t is the transmitted power, $P_r(d)$ is the received power, G_t is the transmitter antenna gain, G_r is the receiver antenna gain, d is the T-R separation distance in meters and λ is the wavelength in meters and L is the system loss factor (≥ 1, for example filter losses, antenna losses, etc…). From now on, without loss in generality, we assume L as unity [10]. The Friis free space equation shows that the received power falls off as the square of the transmitter-receiver separation distance. This implies that the received power decays at a rate of 20 dB/decade with distance.

2.3.1.2 Lognormal Path Loss Propagation Model

A lognormally distributed random variable can be used to characterize the shadowing effects that occur with mean value determined by the transmitter and receiver separation distance.

In this channel model the average path loss for an arbitrary T-R couple,$\overline{L_p}$ (d), is expressed as a function of the distance d by using a path loss exponent, independently of the presence of a direct LOS between the transmitter and the receiver units [10].

$$\bar{L}_p(d)=\bar{L}_p(d_0)+10n\log 10\left(\frac{d}{d_0}\right) \tag{2.7}$$

The order n has the constant value of 2 for LOS links but is usually higher than 2 for multipath channels in cities and urban areas [9].

The measured path loss L(d) at distance can be significantly different from the average value due to, for example, shadowing effects, and in fact, is a Gaussian random variable given by:

$$L(d) = \bar{L}_p(d_0)+10n\log 10\left(\frac{d}{d_0}\right) + X_\delta \tag{2.8}$$

X_δ is a zero-mean Gaussian random variable (in dB) with standard deviation also in dB. The path loss so described is known as log-normal shadowing [6]. d_0 is called the free-space close-in reference distance [10]. The selected value of d_0 must be appropriate for the propagation environment.

2.3.2 Small Scale Fading Channels

The second category of fading channel is Small-scale fading or simply fading which is used to describe the rapid fluctuations of the amplitude, phases, or multipath delays of a radio signal over a short period of time or travel distance, so that large-scale path loss effects may be ignored.

Based on the relation between the signal parameters (bandwidth, symbol period) and the channel parameters (RMS delay spread and Doppler spread), we have four different cases of fading. Due to time dispersive nature we have Flat and Frequency selective fading and due to Doppler spread we have fast and slow fading [12].

Fading might have a time varying or frequency varying attenuating impact on the transmitted signal. Due to the frequency varying and time varying (complex valued) nature of fading, we will denote the attenuating impact in this section by $\overline{H}(t, f)$. In some cases the fading might be only on time varying or frequency varying. We denote the fading by

$\bar{h}(t)=\bar{H}(t,0)$ in the case of time varying fading only, and $\bar{H}(f)=\bar{H}(0,f)$ in case of frequency varying only.

2.3.2.1 Small Fading Effects due to Multipath Time Delay Spread

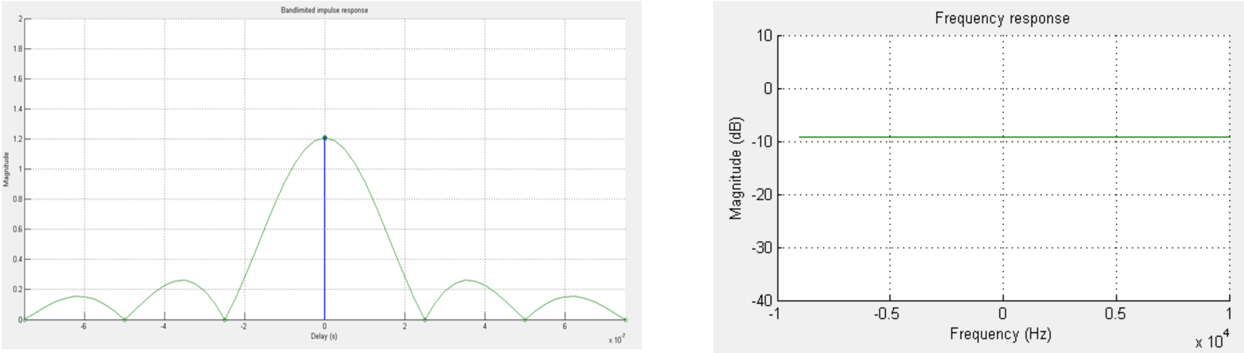
2.3.2.1.1 Flat Fading

Small-scale fading is defined as being flat or non-selective if the received multipath components of a symbol do not extend beyond the symbol's time duration .If the delay of the multipath components with respect to the main component is smaller than the symbol's time duration, a channel is said to be subject to flat fading [2]. In this fading channel ISI is absent; therefore, such a channel has a constant gain and a liner phase response over a bandwidth that is greater than the bandwidth of the transmitted signal

In a flat-fading channel, the spectral characteristics of the transmitted signal are preserved at the receiver, and the channel does not cause any non-linear distortion due to time dispersion. However, the strength of the received signal generally changes slowly in time, due to the slow gain fluctuations caused by multipath. Flat-fading channels are also known as amplitude varying channels, and they are sometimes referred to as narrow band channels, as the bandwidth of the applied signal is narrow with respect to the channel bandwidth [9]. In a flat-fading channel, the following hold true:

$$T_s \gg \sigma_\tau \text{ And } B_s \ll B_c \tag{2.9}$$

Where B_s is the bandwidth of the transmitted signal, $B_c$is the coherence bandwidth of the channel, T_s is the symbol's period, and $\sigma_\tau$is the rms (root mean square) delay spread of the channel. Channel estimation may estimate the phase shifted and attenuated in amplitude without channel equalization since ISI is absent in this channel.



a) Impulse response b) Frequency response

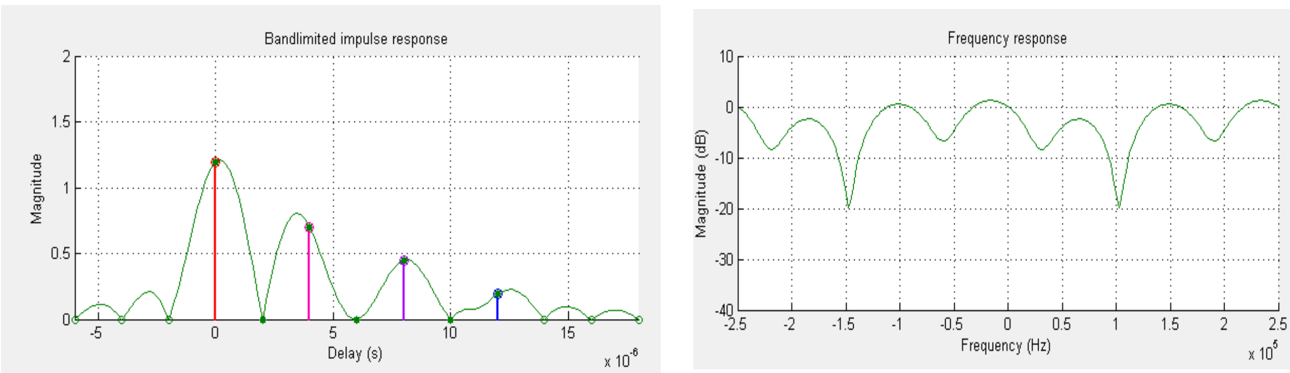
Figure 2.3 Impulse response and frequency response of flat fading channel

2.3.2.1.2 Frequency-Selective Fading

When the signal bandwidth is more than the coherence bandwidth of the mobile radio channel or equivalently the symbols duration of the signal is less than the RMS delay spread [2], frequency selective fading occurs.

$$B_s \gg B_c \text{ and } T_s \ll \sigma_\tau \tag{2.10}$$

At the receiver, we obtain multiple copies of the transmitted signal, all attenuated and delayed in time. The channel introduces inter symbol interference [12]. In frequency selective fading channel equalizers are needed after channel estimation to compensate ISI.



a) Impulse response b) Frequency response

Figure 2.4 Impulse response and frequency response of frequency selective channel

2.3.2.2 Small Scale Fading Effects due to Doppler Spread

Depending on the effect of Doppler Spread fading channel is divided into slow and fast fading fast fading.

2.3.2.2.1 Slow Fading

When the channel impulse response changes rapidly within the symbol duration of the signal fast fading occurs. Since coherence time of the channel is smaller than the symbol period of the transmitted signal it results in Doppler spreading, as a result a signal undergoes frequency dispersion leading to distortion [12].

A signal undergoes fast fading if

$$T_s \gg T_c \text{ and } B_s \gg B_D \qquad (2.11)$$

Where $T_c$the coherence is time and B_D is Doppler spread of the signal.

Slow fading could be caused by some events, for example, shadowing as a large obstruction such as hills, buildings [11].

2.3.2.2.2 Fast Fading

Fast fading occurs when the coherence time is smaller than the delay constraint of the channel. In this condition, the amplitude and phase change imposed by the channel varies considerably over the period of use [11]. Therefore, the signal undergoes fast fading if

$$T_s < T_c \text{ or } B_s < B_d \qquad (2.12)$$

We can observe that the velocity of the user has an important factor in deciding whether the signal experiences fast or slow fading.

This thesis concentrates on the effect of time delay spread flat fading and frequency selectivity.

2.4 Fading Channel Models for Broadcasting Systems

The second generation terrestrial digital television standard DVB-T2 was standardized in 2009. When evaluating the overall performance of such communication systems, different channel models are used to model different reception scenarios [3]. Commonly used channel models are AWGN, Rayleigh, Ricean, RA6 and TU-6. However, since each of the models simulate a specific reception scenario, it is difficult to ascertain the realistic overall performance of the system [3]. As shown in the figure below the received signal is attenuated by the fading coefficient and contaminated by noise.

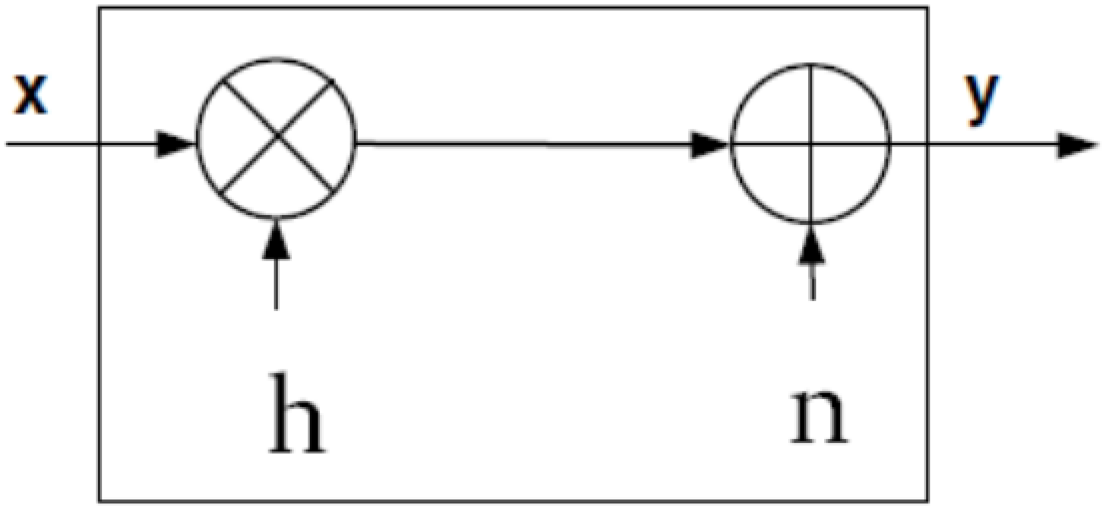


Figure 2.5 Channel model with fading coefficient h and noise n *[10]*

The received signal of this channel model is given by:

$$\mathbf{y = h{*}X + n} \tag{2.13}$$

Where **h** is a new fading coefficient that combines both the large-scale fading and flat fading.

The main job of channel estimation is to estimate the fading coefficients using appropriate channel models. There are many models that describe the phenomenon of small scale fading. Out of these models, AWGN, Rayleigh fading, Ricean fading and COST 207 groups namely TU6 and RA6 will be presented in the following sections

2.4.1 AWGA Channel Model

The simplest radio environment in which a wireless communications system or a local positioning system or proximity detector based on Time of-flight will have to operate is the AWGN environment. AWGN is the commonly used to transmit signal while signals travel from the channel and simulate background noise of channel.

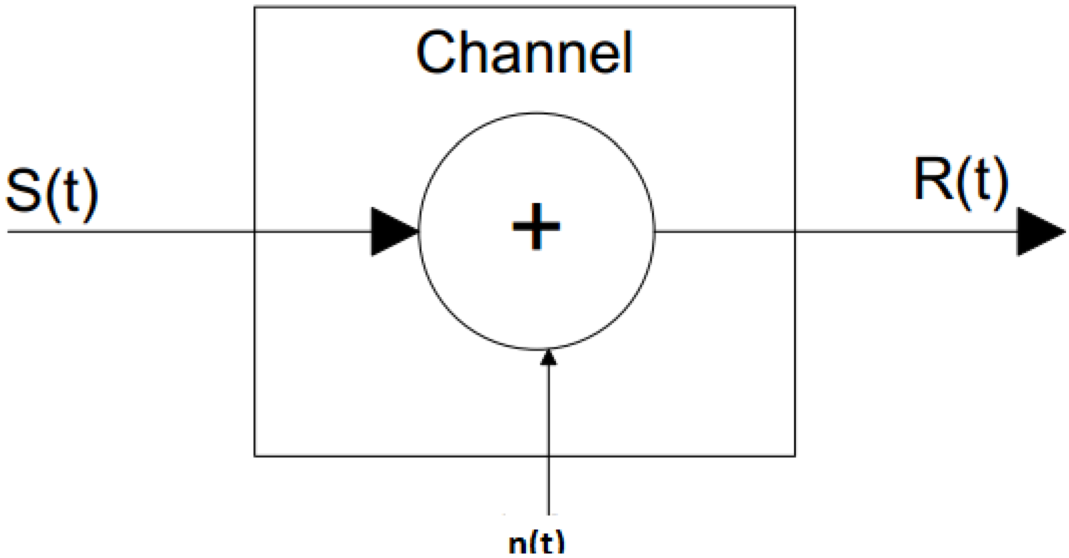


Figure 2.6 AWGN channel model *[10]*

The mathematical expression in received signal

$$R(t) = s(t) + n(t) \tag{2.14}$$

That passed through the AWGN channel where, the transmitted signal s(t) and n(t)is background noise.An AWGN channel adds white Gaussian noise to the signal that passes through it. It is the basic communication channel model and used as a standard channel model. The transmitted signal gets disturbed by a simple additive white Gaussian noise process. The noise n has Gaussian distribution with zero mean and variance of σ^2

2.4.2 Rayleigh Channel Models

The Rayleigh fading is primarily caused by multipath reception. Rayleigh fading is a statistical model for the effect of a propagation environment on a radio signal. It is a reasonable model for troposphere and ionospheres signal propagation as well as the effect of heavily built-up urban environments on radio signals. Fast fading component has Rayleigh density function, if there is no direct path from the transmitter to the receiver. Rayleigh distribution is given by:

$$p(r)=\begin{cases} \frac{r}{\sigma 2}\exp\left\{\frac{-r}{2\sigma 2}\right\} & , 0\leq r < \infty \\ 0 & r < 0 \end{cases} \tag{2.15}$$

Where r is the amplitude of the received signal, and $2\sigma^2$ is the mean power of the multipath signal envelope.

Rayleigh fading is most applicable when there is no line of sight between the transmitter and receiver.

Rayleigh distribution is commonly used to describe the statistical time varying nature of the received envelope of a flat fading signal, or the envelope of an individual multipath component. Envelope of the sum of two quadrature Gaussian noise signals of zero mean obeys a Rayleigh distribution [9]. Flat Rayleigh fading channel can be modeled as shown in Figure 2.7

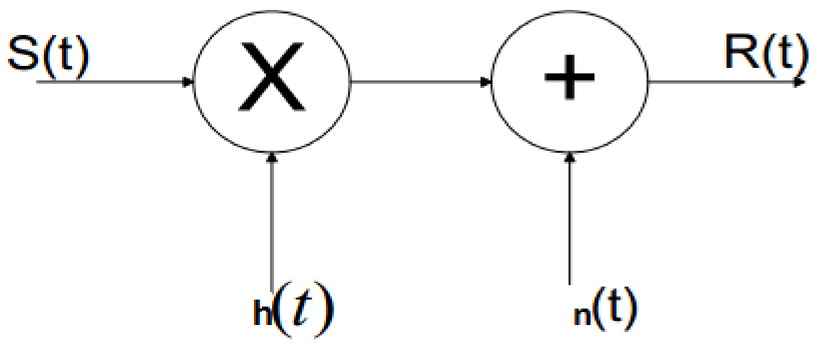


Figure 2.7 Flat Rayleigh fading channel model *[9]*

Where h(t) amplitude fading coefficient.

2.4.3 Racian Channel Models

The Ricean fading model is similar to the Rayleigh fading model, except that in Ricean fading, a strong dominant component is present [9]. This dominant component is a stationary (nonfading) signal and is commonly known as the LOS Component, Rician fading model has a strong dominant component arriving through LOS path, due to which the quadrature Gaussian noise components in the complex Gaussian noise have non zero mean.

If there is a direct path, fast fading component will have Rician density function, which is given by:

$$p_{rician}(r)=\frac{r}{\sigma 2}e^{\frac{-(r^2+A^2)}{2\sigma 2}}I_0\left(\frac{rA}{\sigma 2}\right),\ A\geq 0; r>0 \tag{2.16}$$

Where $I_0\left(\frac{rA}{\sigma 2}\right)=\frac{1}{2\pi}\int_0^{2\pi}\exp\left[\frac{rA\cos\theta}{\sigma 2}\right]d\theta$

Here r is the complex Gaussian vector, σ^2 is the local mean scattered power and A^2 is the power of the dominant component and I_0 is the zero-order modified Bessel function of the first kind.

2.4.4 COST 207 Channel Models

The COST 207 channel models for mobile radio were standardized to enable different communications designers to simulate their systems using a common set of channel models [15]. Four propagation models are defined: (typical rural area –phat6 (RA6), Typical urban area path6 (TU6), bad urban area (BU), and hilly terrain (HT) [15]. Out of these the most commonly used channel models for broadcasting systems are TU6 and RA6 as reference both in DVB-T and DVB-T2 standards.

2.4.4.1 TU6 Channel Models

The TU6 profile reproduces the terrestrial propagation in an urban area. It was originally defined by COST207 [15]. Parameters values fluctuate dynamically, following a Rayleigh law. Each path have Doppler spectrum with Rayleigh-Jakes feature [16]. This channel profile has been proven to present fairly well the general mobile DVB-T reception by several field tests. And again, this channel profile has been used for GSM and DAB tests [16].

TU6 profile consists of 6 paths having wide dispersion in delay and relatively strong power, reproducing an urban environment with NLOS, and consequently, following a Rayleigh model [15]. The following Table 2.2 shows the delay spread and path loss for each paths and Figure 2.8 b show the impulse response of TU6 channel model.

Table 2.1 TU6 channel model parameters *[16]*

Path	Path loss(dB)	Delay (μs)	Doppler Spectrum
1	-3.0	0.0	Rayleigh-Jakes
2	0.0	0.2	Rayleigh-Jakes
3	-2.0	0.5	Rayleigh-Jakes
4	-6.0	1.6	Rayleigh-Jakes
5	-8.0	2.3	Rayleigh-Jakes
6	-10.0	5.0	Rayleigh-Jakes

In DVB-T/H developments the TU6 model was used for modelling mobility of the receiver. TU6 gave reasonable results even though Band widths of GSM and DVB-T/H are very different. One apparent reason for using TU6 also in DVB systems development is that the frequency range is below 1 GHz in both systems [17].

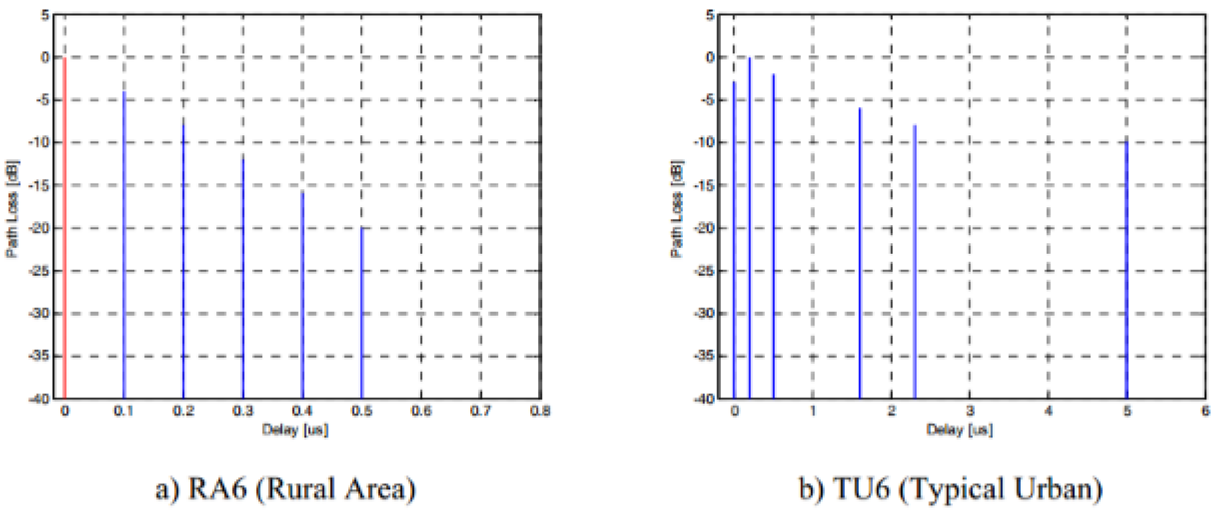


Figure 2.8 Impulse response of the RA6 and TU6 channel models*[16]*

2.4.4.2 RA6 Channel Models

The RA6 channel profile consists of 6 paths having relatively short delay and small power, aiming at reproducing a rural area with LOS and NLOS. Channel RA6 consists of one direct path and five reflected paths and hence, following a Ricean or Rayleigh channel model, respectively.

This profile reproduces the terrestrial propagation in a rural area. It has been defined by COST207 [74] as a "Typical Rural Area" profile and is made of 6 paths having relatively short delay and small power as we can see in Figure 2.8 a. The first part of this channel model has zero delay and attenuation; therefore it is a direct path as we see from Table 2.2).First path has a Doppler spectrum with Rice-Jakes feature. Other paths have Doppler spectrum with Rayleigh-Jakes feature. This channel profile has been used for GSM and DVB tests.

Table 2.2 RA6 (Rural Area) channel model parameters *[16]*

Path	Path loss(dB)	Delay(μs)	Doppler spectrum
1	0.0	0.0	Rice-Jakes
2	-0.4	0.1	Rayleigh-Jakes
3	-8.0	0.2	Rayleigh-Jakes
4	-12.0	0.3	Rayleigh-Jakes
5	-16.0	0.4	Rayleigh-Jakes
6	-20.0	0.5	Rayleigh-Jakes

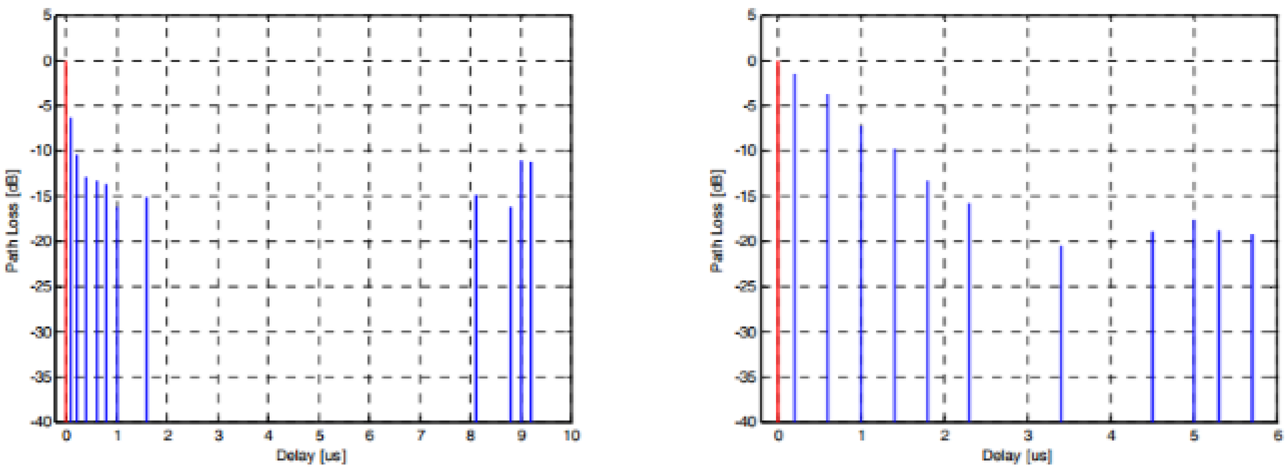
As it can be seen from Table 2.1 and Table 2.2, there are significant differences between the parameters, which defined both channels. In case of RA6 channel, the delay of paths is not higher than 0.5 us, but in the TU6 channel this value is 5us. Moreover, the first path of the RA6 channel is a direct path and the K factor is available and equals to 10. On the other hand, the path losses of echoes are higher in RA6 channels.

2.4.5 Pedestrian outdoor (PO) and Pedestrian Indoor (PI) Channel Models

PO and PI are portable channels. The "portable" means that the device can be easily carried or taken from one point to another. It contains Omni directional antenna and it operates in a nomadic mode (not operated while moving fast). In the context of DVB-T/H, portable antenna reception is defined as the reception at no speed or very low speed (walking speed, approx. 3 km/h) [18].

A PO channel and PI channel were defied for a user velocity of 3 km/h (1.67 Hz of Doppler at 600 MHz) [19]. These channel models are based on measurements in DVB-T/H SFN networks and have paths from two different transmitter locations.

Both channels consist of 12 independent paths, from which the first path is the direct as we can see in Figure. 2.9 a and b. First part has a Doppler spectrum with Rice-Gauss feature. Other paths have Doppler spectrum with Rayleigh-Gauss feature [18].



a) PI (Pedestrian Indoor)   b) PO (Pedestrian Outdoor)

Figure 2.9 Impulse response of PI and Po channel models*[18]*

The PI and PO channel model parameters are summarized in the following Table 2.3

Table 2.3 PI and PO channel model parameters *[18]*

Path	PI		PO		Common for both channels
	Path loss[dB]	Delay[μs]	Path loss[dB]	Delay[μs]	Doppler spread
1	0.0	0.0	0.0	0.0	Rice-Gauss
2	-6.4	0.1	-1.5	0.2	Rayleigh-Gauss
3	-10.4	0.2	-3.8	0.6	Rayleigh-Gauss
4	-13.0	0.4	-7.3	1.0	Rayleigh-Gauss
5	-13.3	0.6	-9.8	1.4	Rayleigh-Gauss
6	-13.7	0.8	-13.3	1.8	Rayleigh-Gauss
7	-16.2	1.0	-15.9	2.3	Rayleigh-Gauss
8	-15.2	1.6	-20.6	3.4	Rayleigh-Gauss
9	-14.9	8.1	-19.0	4.5	Rayleigh-Gauss
10	-16.2	8.8	-17.7	5.0	Rayleigh-Gauss
11	-11.1	9.0	-18.9	5.3	Rayleigh-Gauss
12	-11.2	9.2	-19.3	5.7	Rayleigh-Gauss

For any terrestrial system, non-line of sight components are always present, as the signal is reflected from objects on the way from the transmitter to receiver. In this thesis Rayleigh and TU6 channel models are used for analysis of the proposed channel estimation techniques and algorithms.

Chapter Three

Terrestrial Digital Broadcasting Systems

3.1 Introduction

The success story of radio and television was originally based on terrestrial transmission; although at the end of the twentieth century other distribution forms like cable and satellite had significantly overrun the terrestrial platform in many countries. There are still many regions around the globe where terrestrial broadcasting constitutes the primary means to deliver radio and television programs to the listeners and viewers [20].

For many decades, the quest for better viewing experience has never stopped. Due to dedication of people's efforts, the system went through analog television to digital television broadcasting era. Digital television technology is intended to provide better quality of image as well as sound to the world. This new technology conquered drawbacks of analog television system and has replaced former ones in many developed countries [3]. ATSC standard in North American is developed by United State and further inherited by other countries and DVBT standards in Europe are the main new technologies for this purpose. ATSC standard uses single carrier system as transmission method, whereas DVB standard employed multi-carrier (OFDM) scheme. Single carrier occupies entire bandwidth of channel during transmission which yields higher data rate [3].

In terrestrial broadcasting Multipath propagation affects the performance of broadcasting systems. Its negative influence cannot only be observed in time domain but also when looking at the spectrum of the transmitted signal. The modulation scheme COFDM provides solutions to the problem caused by the multipath propagation effects [21].

Considering television broadcasting, there have not been many revolutionary advances until the end of the late 1990s. However, with the age of digitalization, and after the analog switchover in Europe, it was necessary to develop new terrestrial digital television standards, leading to the DVB-T specification. Currently, with the deployment almost finished in most European countries, a second generation standard DVB-T2 has been developed due to the need for new services that require greater capacity, such as HDTV, and improved coverage [21].

This chapter provides a theoretical background of DVB-T and DVB-T2, as well as a general comparison between them and improvements of DVB-T2 over DVB-T. In addition, special emphasis is done on the improvements introduced by the new standard. The basic transmitter and receiver building blocks of DVB-T2 are then analyzed, specially focusing on pilot and their arrangements in DVB-T2 since this thesis work is piloted based channel estimation techniques for DVB-T2, it will be also considered in Chapter 4 for the channel estimation review, and basic concepts will also be explained.

3.2 Classification of Broadcasting Systems

Generally broadcasting system can be classified as analog and digital broadcasting system. Analog broadcasting is the transmission of sounds and picture through deliberate variation in signal voltage and radio frequencies and has many draw backs compared to digital broadcasting systems. Digital links uses data compression; generally have more bandwidth efficiency than analog links, which allows a content provider more room to provide service, or to provide a higher quality signal than analog links. Our focus will be on the digital part.

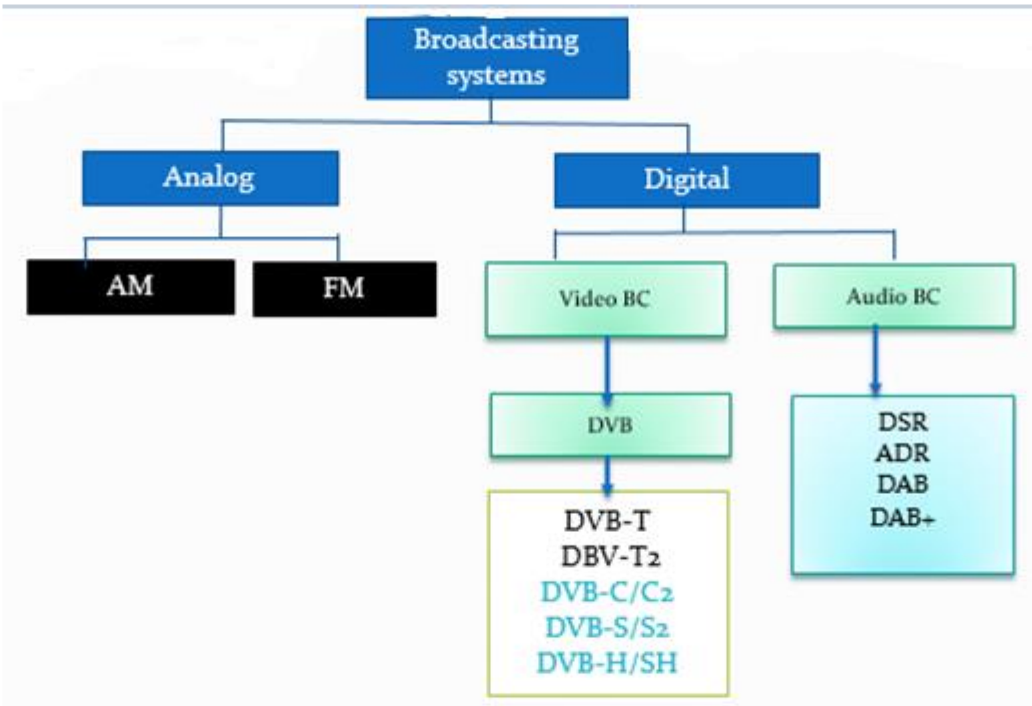


Figure 3.1 Classification of Broadcasting Systems *[22, 23]*

As shown in Figure 3.1 above the digital broadcasting is divided in to video broadcasting and audio broadcasting each classifications contains many standards. Video broadcasting includes DVB set of standards that define digital broadcasting using existing satellite, cable, and terrestrial infrastructures. In the following section we will present DVB standards and their historical evaluation.

3.3 Digital Video Broadcasting

DVB standards family is a set of internationally accepted open standards for digital television. DVB standards are developed by the DVB Project, an international industry consortium with more than 270 members. Its objective is to agree specifications for digital media delivery systems, including broadcasting. It is originally of European origin but now worldwide [3]. DVB is family of standardized technologies designed to facilitate broadcasting of images, sounds and multimedia and to permit a large degree of user interaction. The standard provide for delivery of the data by terrestrial, cable and satellite infrastructures [3]. The various DVB standards include DVB-S for satellites, DVB-C for cables, DVB-T for terrestrial transmission and DVB-H for low power handled terminals.

Among them DVB-T and DVB-H utilize OFDM as modulation scheme. DVB-T/H standards offer several modes of operation that are tailored for large scale single frequency network (SFN) and high mobility reception. DVB-H Digital Video Broadcasting Hand-held is adopted by ETSI as a European Standard of Terrestrial Digital TV for mobile services [24].Comparison of DVB standards is summarized in the following table.

Table 3.1 Comparison of DVB standard families *[25, 23]*

DVB Family	Possible Modulation schemes	Basic characteristics
DVB-T	QPSK,16-QAM,64-QAM	Covers distance of less than 100km, higher power TX, channel is estimation possible /required
DVB-T2	QPSK,16-QAM,64-QAM,256-QAM	
DVB-S	QPSK	Covers long distance~36000km downlink, channel unknown b/c of weather condition(rain, clouds)
DVB-S2	QPSK,8-PSK,16-PSK,32-PSK	
DVB-C	16-QAM32-QAM,64-QAM,128-QAM,256-QAM	Distance~ few km, requires line amplifier (high signal) channel characteristics constant.
DVB-C2	16-QAM32-QAM,64-QAM,128-QAM,256-QAM,1024-QAM,4096-QAM	
DVB-H	QPSK,16-QAM,64-QAM	Mainly designed to enable the provision of mobile broadcasting services.
DVB-SH		

This thesis work focuses on DVB-T and DVB-T2 since Terrestrial transmission path is subject to numerous impacts such as echoes and multipath reception, AWGN and Doppler shift in case of mobile reception that needs channel estimation and channel equalization at the receiver side. The basic principles and basic parameters of DVB-T and DVB-T2 will present in the following sections.

3.3.1 Evolution of DVB Standards

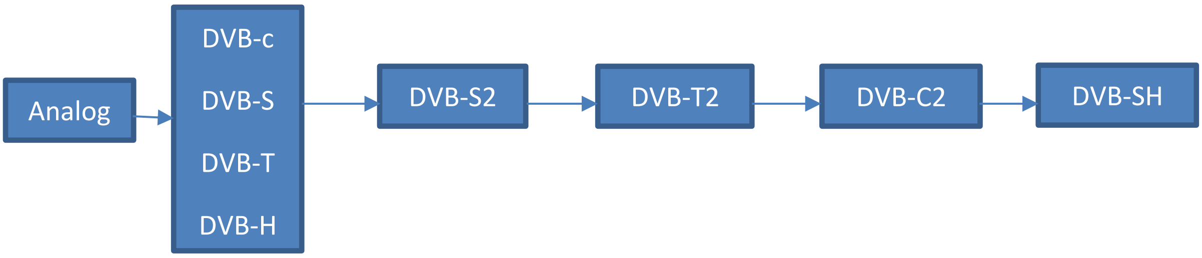


Figure 3.2 Evolution of DVB standards *[26]*

By the mid 1990's, the DVB organization had defined three principal standards, DVB-S, DVB-T and DVB-C for distribution of TV over satellite, terrestrial and cable channels. Ten years later, the DVB organization defined its first DVB-S2 standard [11]. This new satellite standard provided for a 30% increase in transmission capacity compared with DVB-S and was intended primarily for HDTV broadcasting. In DVB-S2, the forward error correction is based on advanced LDPC concatenated with BCH codes [26].

In 2006, it was widely recognized within DVB that, building on the improvements in DVB-S2, similar improvements might be achievable for terrestrial broadcasting and that such improvements could help the terrestrial platform in Europe migrate towards HDTV. Consequently, DVB launched a technical group to define the next-generation, DVB-T2 standard which was eventually published in June 2008 [26]. DVB-T2-based multiplex would be able to deliver at least 30% more capacity than a conventional DVB-T multiplex for the same coverage planning parameters. More than 40 companies have actively participated in the development of the T2 specification through an intense schedule of meetings, telephone conferences, email exchanges followed by validation and verification exercises to test the interoperability of early implementations [26].

Towards the end of this process, DVB started the process of defining a second-generation, DVB-C2 modulation scheme for cable systems. This DVB-C2 standard built on many of the advanced techniques and data structures within the DVB-S2 and DVB-T2 standards leading to a coherent, 'family' of 'DVB-x2' second-generation standards. The DVB-C2 standard was published at the end of 2009 [26]. In the following sections we will focus on DVB-T/T2.

3.4 DVB-T and its Limitations

DVB-T is the first terrestrial digital video broadcasting standard developed for replacing the former analogue systems in many countries around the world. The benefits of digital coding and transmission techniques allow perfect signal recovery in all the serviced areas avoiding the effects of the wireless channel and noise. Considering the physical level of the communications, the digital data sequences, which contain MPEG video, audio and other information streams, are transmitted using COFDM modulation [27]. The information bits are coded, interleaved, mapped to a QAM constellation and grouped into blocks. All the symbols in a block are transmitted simultaneously at different frequency subcarriers using an IFFT operation. The number of IFFT points, which can be either 2048 (2K) or 8192 (8K), determines the transmission mode and the number of the available subcarriers in the transmission bandwidth. Some of these subcarriers are not used to allow for guard frequency bands whereas others are reserved for pilot symbols, which are necessary to acquire the channel information required for signal recovery. Figure 3. 3 shows the main diagram of a DVB-T transmitter. As it can be seen, the data bit stream is scrambled, processed by an outer RS coder, an interleaver and an inner convolutional coder. The first coding stage removes possible error floors at high SNR values, whereas the second reduces the BER at the receiver by including more redundant information depending on the selected CR, which can range from 1/2 to 5/6. The coded information bits are interleaved again in order to allocate consecutive bits to different subcarriers. The resulting information bits are then arranged by blocks, mapped and modulated using OFDM, which involves an IFFT operation and the addition of a cyclic prefix to enable a guard interval that avoids interference between consecutive blocks [27]. The use of coding and interleaving processes over OFDM provides an efficient and robust transmission method in multipath scenarios enabling time and frequency diversity.

Both DVB-T and DVB-T2, use OFDM modulation, which is suitable for terrestrial channels, in which deep selective fades are observed because of multipath propagation and the SFNs operation and use two stages of channel coding [28].

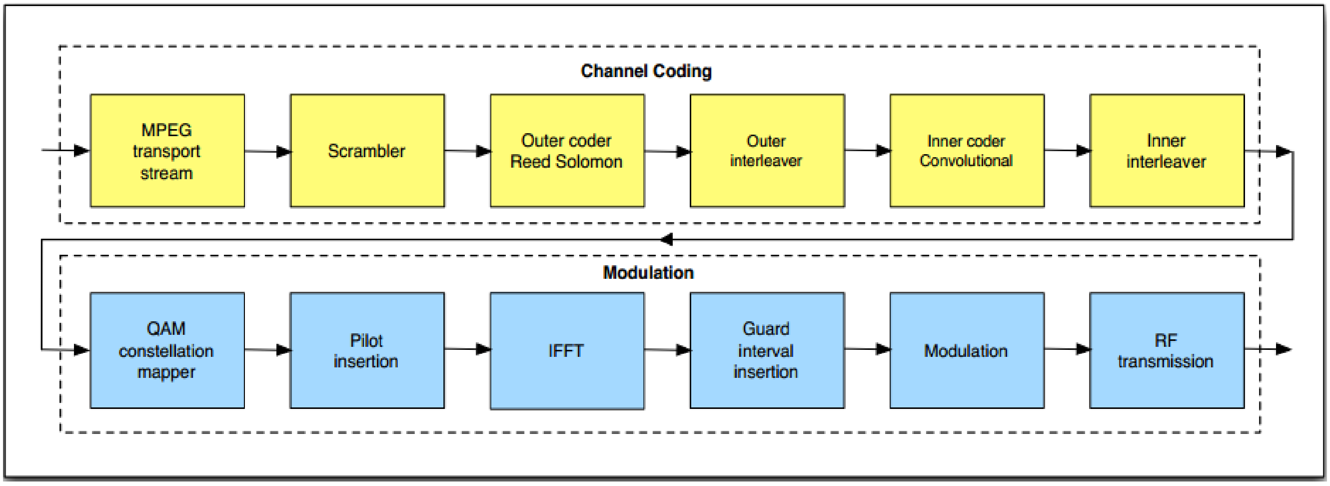


Figure 3.3 Block diagram of a DVB-T transmitter *[21].*

The coded information is again interleaved and thereafter, the input bits from channel coding stage are mapped onto a QAM constellation and modulated using OFDM, which will be briefly explained in the next section.

Despite the many benefits achieved by the deployment of the DVB-T network, its limitations became clear from the beginning. First, the number and bit rates of the transmitted channels are limited in comparison with new wireless transmission techniques. A new standard was soon required to broadcast more channels and HDTV using the same frequency spectrum. Second, a new information system was required to allow more interaction with the user. Third, the DVB-T standard, which had been designed for fixed scenarios, had a very bad performance in mobile or portable environments, so it could not be properly implemented in scenarios such as moving vehicles. Last but not least, the deployment of the DVB-T network has been and still is a true nightmare in SFN scenarios, where interferences between repeaters, which transmit the same information on the same frequency bands, may destroy the received signal avoiding its reception in areas with good reception levels [27].

Considering the new advances in signal processing, modulation and coding, the DVB consortium has published a draft standard named DVB-T2 aiming to extend the capabilities of the aforementioned DVB-T standard. Thought out this chapter we will have more about DVB-T2 frameworks and improvements over DVB-T.

3.5 The New DVB-T2 Standard

DVB-T2 is the second-generation terrestrial transmission system for digital television broadcasting. DVB-T2 is not originally designed to replace DVB-T in short term, and both will coexist in the market for services of different needs in the future. The DVB organization has defined a series of business requirements as the development framework for DVB-T2.

The motivation for deployment of DVB-T2 is to support HDTV services as effectively and efficiently as possible by providing an additional 30% capacity whilst using the existing infrastructure at both the transmitter and the receiver [29]. Based on recent research results and a set of commercial requirements, the DVB consortium concluded that there were suitable technologies which could provide increased capacity and robustness in the terrestrial environment, mainly for HDTV transmission. Therefore, a new standard named DVB-T2 has been designed primarily for fixed receptors, although it must allow for some mobility, with the same spectrum characteristics as DVB-T [27].

DVB-T2 builds on the technologies used as part DVB-T and adopts many sophisticated technologies, such as Physical Layer Pipe technology and low-density parity check codes to support highly flexible and reliable transmission [29]. The coding algorithms, based on the combination of LDPC and BCH codes, offer excellent performance resulting in a very robust signal reception. LDPC-based FEC techniques can offer a significant improvement compared with the convolutional error correcting scheme used in DVB-T [27].

Regarding the modulation, DVB-T2 uses the same OFDM technique as DVB-T. The new specification offers a large set of modulation parameters by combining different numbers of carriers and guard interval lengths, making it a very flexible standard as it is shown in Table 3.1 Furthermore, the highest constellation size has been increased to 256 symbols (256QAM).

3.6 Transmitter Blocks of DVB-T2

The generic block diagram of a DVB-T2 transmitter is depicted in Figure 3.4. As it can be seen, it is divided into four main parts; input processing, bit interleaved coded modulation (BICM), frame builder and OFDM generation. The system input(s) may be one or more MPEG-2 Transport Stream(s) and/or one or more Generic Stream(s). The Input Pre-Processor, which is not part of the T2 system, may include a Service splitter or de-multiplexer for TS for separating the services into the T2 system inputs, which are one or more logical data streams.

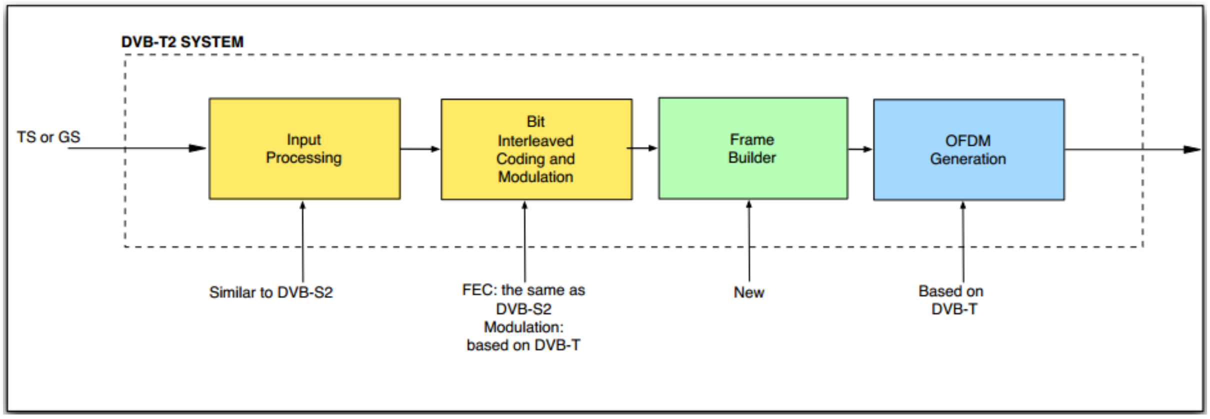


Figure 3.4 High level DVB-T2 block diagram *[21]*

In Figure 3.5 the system block diagram at the transmitter side is shown in detail, being the reverse process performed at the receiver side. In DVB-T2 the input processing and FEC mechanisms are similar to DVB-S2 and OFDM generation and modulation are developed based on DVB-T Principle. Each blocks of DVB-T2 transmitter will be briefly described in the following subsections.

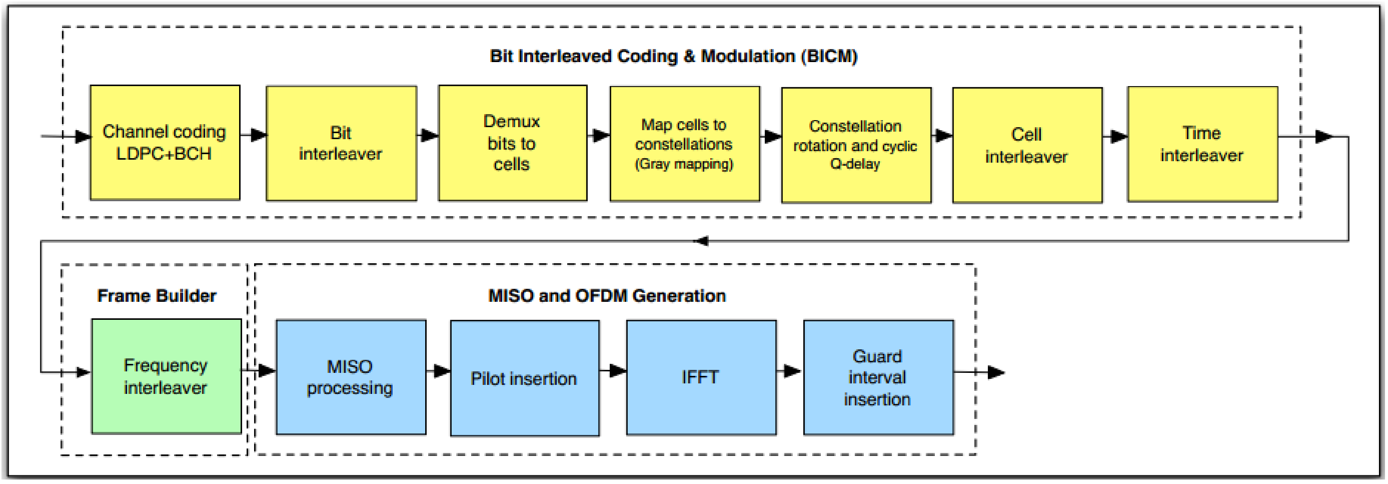


Figure 3.5 Block diagram of a DVB-T2 transmitter *[21]*

3.6.1 Input Processing

The input to the DVB-T2 system shall consist of one or more logical data streams. One logical data stream is carried by one Physical Layer Pipe. The mode adaptation modules, which operate separately on the contents of each PLP, slice the input data stream into data fields which, after stream adaptation, will form baseband frames (BBFRAMEs).

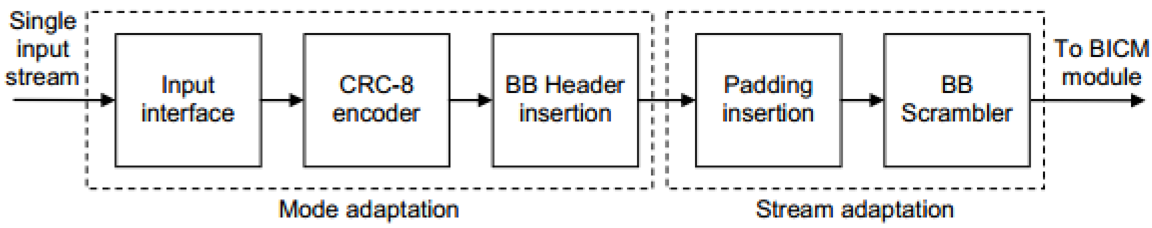


Figure 3.6 system block diagram of input processing module for input mode A (single PLP) *[30]*

Mode A when a single PLP is used (Mode A), there will be a constant number of FEC blocks per Interleaving Frame and each Interleaving Frame will correspond to one T2 frame. This in turn will correspond to a fixed number of input bits, and the scheduler need only count these bits and map them into the BBFRAMEs whereas Mode B When multiple PLPs are used (Mode B), with potentially variable bit-rates in each PLP, the situation is more complicated [30].

3.6.1.1 Mode Adaptation

At the beginning of the DVB-T2 system the input interface may have preprocessor that takes TS or GS as inputs. Afterwards the service splitter inside the input preprocessor may split the services of TS into logical data streams, which are then carried by individual PLP. Later mode adaptation module individually works on the contents of each PLP by slicing the data stream in to data fields and add baseband header at the start of each field [31].

As we can see in Figure 3.6 the mode adaptation module comprises the input interface, followed by three optional sub-systems (the input stream synchronizer, null packet deletion and the CRC-8 encoder) and then finishes by slicing the incoming data stream into data fields and inserting the baseband header (BBHEADER) at the start of each data field [30]. Each of these sub-systems is described in the following clauses

Input Interface

The input interface subsystem shall map the input into internal logical-bit format. The first received bit will be indicated as the Most Significant Bit (MSB). Input interfacing is applied separately for each single PLP. The Input Interface shall read a data field, composed of DFL bits (Data Field Length), where: $0 < \mathrm{DFL} < (\mathrm{k}_{\mathrm{bch}-80)}$ where $\mathrm{k}_{\mathrm{bch}}$ is the number of bits protected by the BCH and LDPC codes [32]. The maximum value of DFL depends on the chosen LDPC code, carrying a protected payload of Kbch bits. The 10-byte (80 bits) BBHEADER is appended to the front of the data field, and is also protected by the BCH and LDPC codes [30].

CRC-8 encoding

CRC-8 is applied for error detection at UP level (Normal Mode and packetized streams only). When applicable, the UPL-8 bits of the UP (after sync-byte removal, when applicable) shall be processed by the systematic 8-bit CRC-8 encoder. The computed CRC-8 shall be appended after the user packet [30].

Baseband Header (BBHEADER) Insertion

A fixed length BBHEADER of 10 bytes shall be inserted in front of the baseband data field in order to describe the format of the data field. The BBHEADER shall take one of two forms NM and HEM [30].

3.6.1.2 Stream adaptation

Stream adaptation module takes a baseband header followed by a data field and make a baseband frame. It comprise of three sub-modules: scheduler, Padding insertion and BB scrambling. Since our focus is on mode A we will skip the scheduling part [33].

Padding insertion

When the baseband frame has insufficient data to fill it up or has the requirement to have an integer number of user packets, then padding is used to append zero bits after the data field to fill the frame [31].

DVB-T2 offers three mechanisms for padding: Null packets (without NPD), BBFRAME padding and Dummy cells padding [32]. Although null-packet deletion is available as an option, the system can carry some null packets in their entirety. The two techniques may be mixed, with some null packets transmitted and others deleted and re-inserted [32].Finally, there will almost always be dummy cells inserted at the end of the T2-frame: as the T2-frame is very unlikely to contain a whole number of FEC blocks, the remaining space at the end of the T2-frame will be filled by dummy cells. The first two methods are therefore preferable in most applications. The choice between them is likely to depend on the nature and source of the input stream [32].

BB scrambling

The main purpose of baseband scrambler is to ensure that the complete baseband frame is randomized. In addition, there must be synchronization between the randomization sequence and the baseband frame [31].

3.6.2 Bit-Interleaved Coding and Modulation

BICM module takes a baseband frame as input and produces an output for the next frame mapper. To carry out this task, the BICM shall perform outer coding (BCH), Inner Coding (LDPC), bit interleaver and demux for mapping bits to QAM constellation cells. As we can see in the figure below.

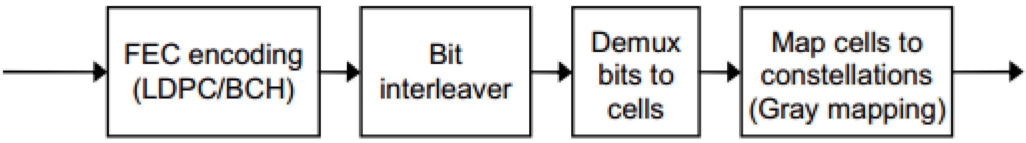


Figure 3.7 DVB-T2 BICM scheme*[32, 30]*

3.6.2.1 FEC Channel Encoding

The purpose of the channel encoder is to introduce redundancy in the information sequence that can be used in the receiver to overcome the effects of noise and interference, therefore improving the transmission reliability. DVB-T2 has to face the challenge of increasing the capacity without increasing the power and bandwidth consumption, in order to be compatible with its predecessor [21]. This enhanced capacity is mainly obtained from the inclusion of channel codes with higher correction error capability. DVB-T2 uses a powerful coding scheme based on the combination of LDPC and BCH error correcting codes [21].

LDPC is inner code which works well only for randomly distributed bit errors used to avoid regular patterns of errors and bursts of errors. For the LDPC code, DVB-T2 includes two choices of code length (Nldpc = 64 800 for normal blocks, and 16 200 for short blocks) and 6 choices of code rate 1/2, 3/5, 2/3, 3/4, 4/5, and 5/6 are defined for both normal and short codes [32].

BCH is outer code is included as an insurance against **u**nwanted error floors at high C/N ratios, and is exactly the same as DVB-S2 [32].

3.6.2.2 Bit Interleaver

LDPC codes in DVB-T2 are irregular LDPC codes and the error-protection level of each code bit is not uniform, but depends on the column weight of the parity-check matrix. The protection level among bits in a multi-level-constellation symbol is not uniform, either. The performance of an LDPC code with multi-level constellation depends highly on the correspondence between code bits and constellation bits. For the purpose of controlling this correspondence, a bit interleave and demuxer are required [32]. The bit interleaver and de-mux were newly designed for 16-QAM/64-QAM/256-QAM constellations [32].

The bit interleaver is a block interleave applied to each LDPC code word. In DVB-T2, a bit interleaver having N_c = 2m columns is used for the 2^m-QAM constellation - except for 256-QAM with the short code, which uses m = 8 columns. The de-mux part de-multiplexes 2m bits in the same row to compose two constellation symbols (except 256-QAM) [32].

3.6.2.3 Demux bits to Cells

After bit interleaving, this sub-module de-multiplexes all the bits into parallel cells so that later these cells can be mapped in to constellation points

3.6.2.4 Constellation Mapping

The constellation mapping for QPSK, 16-QAM, 64-QAM is the same as for DVB-T. In addition, Gray mapping of 256-QAM is added in DVB-T2 [32].

3.6.3 Interleavers

Interleaving is used to spread the content in the time/frequency plane in such a way to combat signal fading by spreading out adjacent data over multiple blocks and sub carriers. This way, error bursts do not affect consecutive bits and can be easier corrected by the channel decoder [21].

There are four different interleavers in the DVB-T2 specification, occurring in the following order in the modulation chain: bit interleaver, cell interleaver, time interleaver and frequency interleaver.

The Bit Interleaver interleaves the code-bits within an LDPC code word in order to avoid undesirable interactions between the bits carried by the same cell and the structure present in the LDPC code [22].

The Cell Interleaver applies a pseudo-random permutation to the cells within a FEC block, with the permutation varying from one FEC block to the next. This disrupts the structured nature of the time interleaver and prevents interactions with the structure of the LDPC code. For example, a burst of erasures caused by a channel will appear in random positions in the FEC code word instead of being regularly spaced, which might interact with the regular structure of the code [22].

The Time Interleaver provides the bulk of the interleaving, spreading the cells of each FEC block over many symbols and potentially over several T2-frames. Time-interleaving ensures that several consecutive bits are transmitted separately [34]. This offers protection against impulsive interference as well as time-varying channels [22].The interleaving time may range from a few milliseconds up to some seconds [34].

Finally, the Frequency Interleaver which is a pseudo-random block interleaver operating on OFDM symbols, that ensures that symbols that have suffered deep fading are spread out across frequency sub carriers [22].

The most significant step from DVB-T to DVB-T2 is the introduction of time interleaving, typically over 70ms, to provide protection against impulsive noise and short time-selective fading [26].Time interleaving also improves the Doppler behavior of the system [34].

3.6.4 Frame Builder

The function of the frame builder is to assemble the cells produced by the time interleavers for each of the PLPs and the cells of the modulated L1 signaling data into arrays of active OFDM cells corresponding to each of the OFDM symbols which make up the overall frame structure. The frame builder operates according to the dynamic information produced by the configuration of the frame structure [32].

At the top level the frame structure consists of the super frame which is divided in to multiple T2 frames, and these T2 frames are further divided in to OFDM symbols. The super frame structure may also include FEF as shown in Figure 3.8 below.

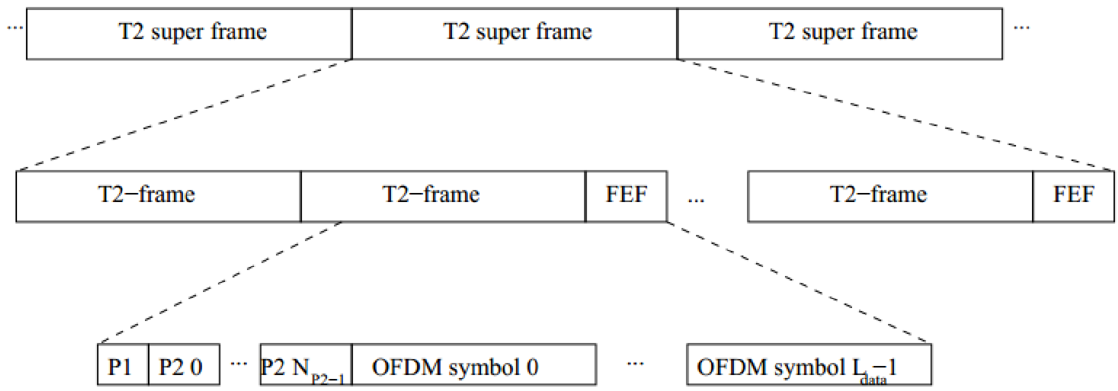


Figure 3.8 DVB-T2 frame structure *[4]*

3.6.4.1 Super Frames

Super frames can carry a configurable number of T2 frames and FEF parts. The maximum number of T2 frames in a super frame is 255, and all the frames within a super frame must have the same length, which can be up to 250ms. Hence, the maximum length of a super frame is 63.75 s, if FEFs are not used. The pattern insertion scheme of FEFs is configurable in each super frame. The maximum number of FEFs in a super frame is 255, reached when one FEF is inserted after every T2 frame [19].

3.6.4.2 Frames

T2 frames are intended to carry DVB-T2 services and related signaling by means of OFDM symbols. They consist of one P1 symbol, one or several P2 symbols (depending on the FFT size), and a configurable number of data symbols [19].

3.6.4.3 OFDM Symbols

The initial symbol of each T2 frame and FEF is a preamble OFDM symbol known as P1. The P1 symbol has a constant 1K FFT size, with guard intervals at both ends. It allows fast identification in the initial scan for detecting the presence of DVB-T2 signals on a given frequency. It carries some basic transmission parameters, such as the frame type (e.g., T2, T2-Lite, or NGH), and it enables the reception of the P2 symbol(s). The subsequent OFDM symbols of a T2 frame are also preamble symbols called P2 symbols. The P2 symbols have the same FFT size as the data OFDM symbols, but have an increased number of pilots. The number of P2 symbols depends on the FFT size used (e.g., two symbols for 8K FFT) [19].

3.6.4.4 Future Extension Frames in DVB-T2

FEFs provide a method for expanding the DVB-T2 standard by means and ways not known at the time of writing the original specification. The use of FEFs is optional. They are inserted, when needed, between T2 frames in such a way that they enable a flexible mixing of services within a single multiplex in a time-division manner. FEFs may also be empty or contain no data [19].

3.6.5 OFDM Generation

The function of the OFDM generation module is to insert pilot in to the cells the frame builder for the OFDM modulation and generate the time domain baseband data signal for the transmission. It then insert the guard interval and, if needed, applies PAPR reduction processing to create the complete T2 signal. Another optional initial stage, known as MISO processing allows the initial frequency domain to be processes by a modified Alamouti encoding [35]. The block diagram of the OFDM generation module is shown in the figure 3.9 below. We will see the basic blocks of OFDM generation module leaving aside the optional one

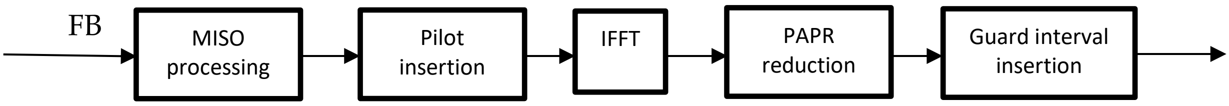


Figure 3.9 block diagram of the OFDM generation module *[36]*

3.6.5.1 Pilot Insertion

Pilots will be inserted in to encoded date cells which has known amplitude and phase that are used by the receiver to compensate/equalize for channel impairments as the channel changes in frequency and in time. The pilots can be used for frame synchronization, frequency synchronization, time synchronization, channel estimation, transmission mode identification and can also be used to follow the phase noise [37].

3.6.5.2 IFFT Insertion

Since OFDM is used as the modulation technique in DVB-T2 system, so this sub-module performs IFFT in the transmitter to convert signal data from frequency domain to time domain.

3.6.5.3 PAPR Reduction

OFDM signals suffer from high Peak to Average Power Ratio (PAPR). DVB-T2 provides two complementary techniques for reduction of PAPR ACE and Tone Reservation. One or both techniques may be used simultaneously.

ACE provides significant reduction in PAPR for lower order constellations (especially QPSK) while tone reservation provides greater benefit for higher order constellations [32]. It is possible to use both ACE and TR simultaneously; however, depending on the constellation, picking the better technique based on the constellation will provide most of the benefit. Both techniques come at a cost: ACE increases the noise level that the receiver sees while TR decreases throughput [32].

3.6.5.4 Guard Interval Insertion

Guard interval is the cyclic continuation of the useful data that is inserted at the begging of the useful part of every OFDM symbol. This sub-module inserts one of the guard interval fractions among the seven different choices available for each FFT size in DVB-T2 [31].

3.6.6 Modes of Transmission in DVB-T2

Traditionally, all communication systems require some core elements in order to operate and propagate the signal. Once all procedures that are necessary are completed, a mode of transmission is selected to feed the signal into the channel. The DVB-T standard offered only the option of the SISO transmission of signals using a transmitting antenna on the provider's end and a receiving antenna on the receiver of the signal end [38].

DVB-T2 standard steps forward to implement additional modes of transmission. The system output that is typically transmitted over a radio frequency channel can support multiple output signals over multiple antennas. Thus, the new standard implements SISO, SIMO, MISO and MIMO transmission modes.

Apart from the SISO mode, the other types implement the approach of spatial diversity by transmitting multiple copies of the same data sequence to combat the channel's performance degradation due to inter-symbol interferences and multipath fading. This technique ensures that the copies data sequences will arrive at the receiver with the lowest amount of fading that is possible.

Spatial diversity on the other hand requires additional components and adds computational complexity to the system that is counter-measured by the use of COFDM modulation. [20]

The DVB-T2 model developed for support all the above modes, but in the context of this thesis we will have brief presentation of the operation SISO with relating to channel estimation in Chapter four of this thesis work.

3.6.7 Choice of Parameters in DVB-T2

There are a large number of ways of configuring a DVB-T2 system. In this subsection we will discuss the factors affecting the choice of each of the main parameters in turn.

Choice of OFDM FFT size

The choice of the OFDM size depends on the application and expected channel characteristics, so T2 gives the content distributors a wide range of possible parameters to adapt to the desired behavior and expected channel suppositions.

The results of choosing a certain FFT size are well known: an increased FFT size will give a greater delay tolerance for the same fractional guard interval, allowing larger SFNs to be constructed. Alternatively, a larger FFT size allows the same delay tolerance to be achieved with a smaller overhead due to the guard interval. A larger FFT size implies a longer symbol duration, which means that the guard interval fraction is smaller for a given guard interval duration in time.

For delivering high-bit-rate services to fixed, rooftop antennas, in the UHF IV/V bands, or lower bands, the 32 K FFT mode is recommended. In this situation the time variations in the channel are minimized, and 32 K offers the very highest bit rates achievable using DVB-T2 [22].For mobile reception in UHF band IV/V, or higher bands, smaller FFT sizes should be used. For mobile use in VHF band III (about 200 MHz) it should be noted that roughly the same Doppler performance should be expected using the 32 K mode as is possible using the 8 K mode at 800 MHz, so 32 K may be an option even at VHF using 7 MHz RF bandwidth. The performance in time-varying channels can also be affected by the choice of pilot pattern [22].

Choice of Guard interval

DVB-T2 offers a wide range of possible guard intervals in order to support a range of broadcasters' needs. We should distinguish between guard-interval duration TG; and guard-interval fraction GIF= TG/ TU. The former has the dimensions of time and the latter is dimensionless. The simplest view is to treat the guard interval as a hard limit to the channel extent that can be tolerated by the system. Assuming the channel extent for a particular broadcast scenario is known, it is then a simple matter to choose a TG that suffices to match or exceed it. Note that this choice will also require consideration of the FFT size as well. The greatest capacity is given by minimizing the GIF, which thus implies maximizing TU. However, there are other constraints on the choice of FFT, concerning the degree of Doppler effects to be expected in the scenario of interest. These may set a limit to the FFT size (and thus TU) that can be chosen [22].

If the same Guard Interval fraction is maintained while the FFT-size of the OFDM symbol is increased, this results in a longer guard interval, and thus a longer echo travel distance. Longer echo travel distances are helpful to build larger Single Frequency Networks, as the distance between two senders of that SFN can be larger [39]. The other way, the Guard Interval fraction can be decreased with increased FFT-size the guard-interval duration will stay the same - to achieve the same maximal echo travel distance for ISI-free reception. Reducing the Guard Interval fraction leads to a shorter. OFDM symbol duration without affecting the useful part of the symbol and the shorter OFDM symbols result in a higher throughput. The overhead between Guard Interval fractions to useful symbol part is reduced [39].

Choice of pilot pattern

Choice of scattered pilot pattern depends on guard interval and FFT size. There is a tradeoff between the bandwidth efficiency and channel estimation accuracy regards the number of pilot symbols. The number of pilot symbols will reduce the bandwidth efficiency and channel estimation accuracy, it is very important that pilot pattern should be chosen carefully in OFDM system [29].

The operator should choose a pilot pattern considering the expected channel for the type of usage that it is desired to support, and being aware of the trade-off between capacity and performance.

Choice of frame length

The DVB-T2-frame length, i.e. the number of symbols L_F in a T2-frame, is a configurable parameter and can be chosen as required by the broadcaster or network operator. There are a number of other factors affecting choice of frame length:

- A longer value for frame length generally decreases the percentage overhead associated with the preamble symbols P1, the L1 signaling and the higher density of pilots in the P2, thus increasing the total bit-rate (but see below).
- A shorter value for frame length means that the P1 and P2 occur more frequently, allowing faster lock-up and service acquisition.
- Longer frames may be used support longer time-interleaving depths.
- Shorter frames may be used to achieve higher peak bit-rates without using multiple time-interleaving blocks [22].

The exact choice of frame length will depend on the selections made for several of the other parameters, including FFT size, Guard interval, the use of extended-carrier mode and the combinations of different types of PLPs [22]

Choice of time interleaving parameters

The choice of number of TI blocks per Interleaving Frame depends on two factors. Increasing the number of TI blocks for a given frame duration reduces the interleaving time, so reducing the time diversity and therefore the system's resistance to impulsive interference and fast time-varying channels. On the other hand, increasing the number of TI blocks increases the maximum data rate for a PLP, since the maximum number of cells in a TI block is fixed [22].

Choice of code rate, block length and constellation

The principles affecting choice of code-rate and constellation are similar to those in DVB-T**.** Higher code-rates and higher-order constellations both give greater bit rates but require higher signal-to-noise ratios. However, it should be noted that the performance of LDPC codes is significantly better than convolutional codes used in DVB-T [22]. Consequently, the SNR requirement for a given LDPC code-rate is lower than for a convolutional code with the same code rate. Conversely, at a given SNR, a higher code rate can be used and therefore a higher data rate achieved [22].

The performance of the short codes is some tenths of a dB worse than normal codes. However, they can be used for low-bit-rate applications requiring shorter latency. Alternatively, if long blocks are used for low-data-rate services, a user could choose either multi-frame interleaving (giving greater time diversity) or frame skipping (giving potential for power-saving); both options would entail a greater latency [22].

3.7 Improvements of DVB-T2 over DVB-T

DVB-T & DVB-T2 standards are similar to each other as both use COFDM modulation but DVB-T2 is distinguished by a number of technical enhancements. The primary requirements of DVB-T2 were to increase the data carrying capacity by up to 30% over DVB-T and also to improve the flexibility of operation and robustness of reception [40]. Some of improvements are not included totally in DVB-T and some of them are improved from DVB-T. In the DVB-T there are no mechanisms protecting from high PAPR, which is a drawback of this standard.

In terms of Signal bandwidth nominal bandwidth, which is the frequency difference between the boundary sub-carriers K_{max} and K_{min} for DVB-T signals equals 7.61 MHz in 8 MHz channels and 6.66 MHz in 7 MHz channels [28]. But in DVB-T2 the set of possible bandwidth options is wider: 1.7 MHz, 5 MHz, 6 MHz, 7 MHz, 8 MHz and 10 MHz Focusing on the most popular 8 MHz bandwidth, the nominal bandwidth of so called normal carrier mode of DVB-T2 signal is the same as in the DVB-T case, i.e. 7.61 MHz for all FFT modes [28]

In summary the following shows the comparison between DVB-T and DVB-T2 including improvements of DVB-T2 over DVB-T

Table 3.2 comparison between DVB-T and DVB-T2 *[40, 26, 41]*

Parameters	DVB-T	DVB-T2
FEC and code rates	Convolutional Coding+Reed Solomon 1/2, 2/3, 3/4, 5/6, 7/8	LDPC + BCH 1/2, 3/5, 2/3, 3/4, 4/5, 5/6
Modulation	QPSK, 16QAM, 64QAM	QPSK, 16QAM, 64QAM and 256QAM
Rotated constellation Mode	N/A	Rotated or None rotated modes are possible
Interleavers	Outer and inner interleavers	Bit, cell, time and frequency interleavers
Guard Interval	1/4, 1/8, 1/16, 1/32	1/4, 19/128,, 1/8, 19/256, 1/16, 1/32, 1/128
FFT Size	2k, 8k	1k, 2k, 4k, 8k, 16k, 32k
Scattered Pilots	8% of total	1%, 2%, 4%, 8% of total
Pilot Patterns	N/A	8 Patterns Available
Continual Pilots	2.6% of total	0.35% of total
PAPR reduction	Not present	Present
FEF	Not present	Present
Bandwidth	6, 7, 8 MHz	1.7, 5, 6, 7, 8, 10 MHz
Typical data rate (UK)	24 Mbit/s	40 Mbit/s

3.8 Receiver Blocks of DVB-T2

Some parts of a receiver correspond directly to a block or feature in the modulator, and inverse the operation of the transmitters. In some cases an algorithm exists which is more efficient in terms of memory or processing on the receiver side, and some very short brief will be given in this subsection. Other parts of the receiver, such as synchronization, channel estimation and channel equalization have no counterpart in the transmitter part. The main part of this thesis work i.e. channel estimation will be present in the next chapter.

The receiver comprises the above blocks in Figure 3.11 Most simply implement the reverse of the corresponding transmitter module and will not be discussed further:

Three reverse of transmitter blocks are perform before channel estimation. The first block is P1 extraction used to removes and discards the P1 preambles. This module also removes and discards any FEF parts. No attempt is made to decode either. The guard interval inserted in the OFDM generation module should remove and discard in the GI removal block of the receiver. FFF Performs the inverse FFT on the active symbols. The considerations for the use of the FFT algorithm in a receiver are essentially the same as those for the use of the IFFT in the modulator. We need to convert the time domain signal to frequency domain at the receiver.

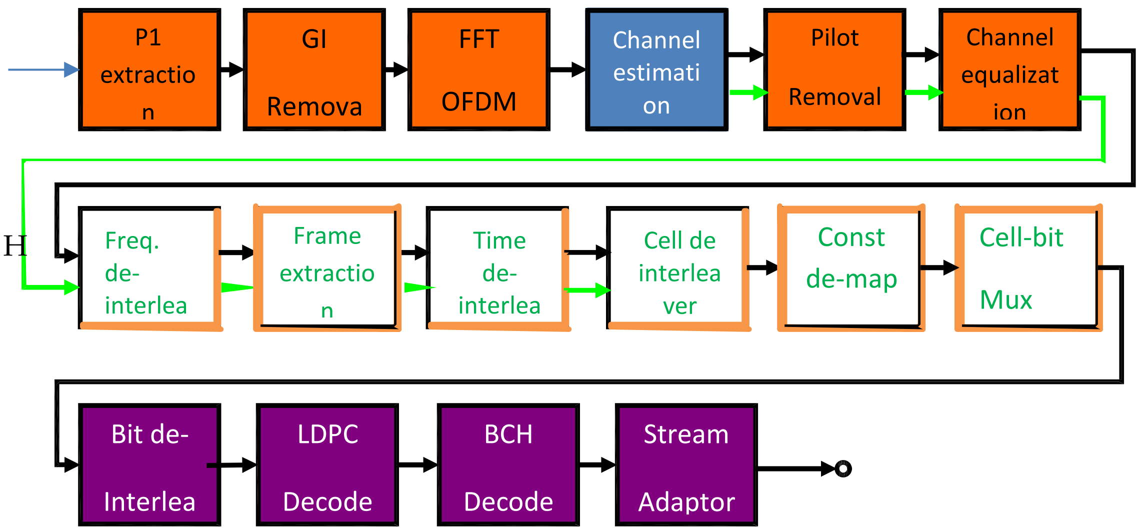


Figure 3.10 DVB-T2 Receiver block *[26, 42]*

After channel estimation is completed pilot symbols should be removed at the Pilot removal block since pilots doesn't carry any useful information [42].

The operation of De-interleaving, demodulation, de-mapping and decoding are used to inverse of the operation that is added in the transmitter. Frame extraction also reverse of frame builder. As discussed in 3.6.4, the receiver currently only processes one PLP, so this module only extracts the cells of this one PLP to feed to subsequent modules.

The compensation of the channel effect is also done at the receiver side by applying synchronization, channel estimation and channel equalization. The main focus of our thesis work is implementation of channel estimation technique pilot assisted. The detail of channel estimation techniques and algorithms for DVB-T2 will be given in the next chapter.

Chapter Four

Channel Estimation

4.1 Introduction

A channel is a medium, which transfer data or information from transmitter to receiver. The feature of any physical medium is that , the transmitted signal is corrupted in various ways by frequency and phase distortion, inter symbol interference, thermal noise etc. and the receiver receives corrupted signal [43]. Estimation is calculation, characterizing the effect of the physical channel on the input sequence. We can say a channel is well estimated when its error minimization criteria is satisfied [44].

Channel estimation algorithm explains the behavior of the channel and allows the receiver to approximate the impulse response of the channel [44]. The estimated channel information is used by equalizers so that the fading effect and/or co-channel interference can be removed and the original transmitted signal can be restored [3].

In this chapter, the three most common channel estimation techniques pilot-based, blind and semi-blind channel estimation will be reviewed, with special emphasis on pilot-based. Taking the pilot type and arrangement presented in Chapter three used as reference, different channel estimation and interpolation techniques are analyzed for DVB-T2 systems, with slow mobility and without mobility. Furthermore, the channel estimation algorithms used to implement each channel estimation techniques will be studied. The pilot-based channel estimation algorithms like LS, MMSE, LMS and RLS are compared based on complexity, speed of convergence and suability at the end of this chapter and LS and MMSE algorithms are implemented for extracting of the channel coefficients for DVB-T2.

4.2 Classification of Channel Estimation Techniques and Algorithms

There are different perspectives on how to categorize channel estimation techniques. They are (i) based on the historical development of estimators and receivers; the techniques which are classified in this category are: (a) The traditional channel estimation technique, Channel Frequency Response (CFR), which consists of Pilot Aided Channel Estimation (PA-CE), Decision Directed (DD-CE), Super Imposed (SP-CE) and Transform Domain (b) Parametric Model (PM) and (c) Iterative Channel Estimation:- e.g Turbo receiver channel estimation and Factor graph channel estimation [45], (ii) based on the insertion or use of training signals such as blind, non-blind and semi-blind [5]. We will focus on the second classification perspectives which is based on the requirement of training sequence or not.

Figure 4.1 below shows the classification of the channel estimation techniques and algorithms based on the insertion or use of training signals such as blind, non-blind and semi-blind in block diagram form.

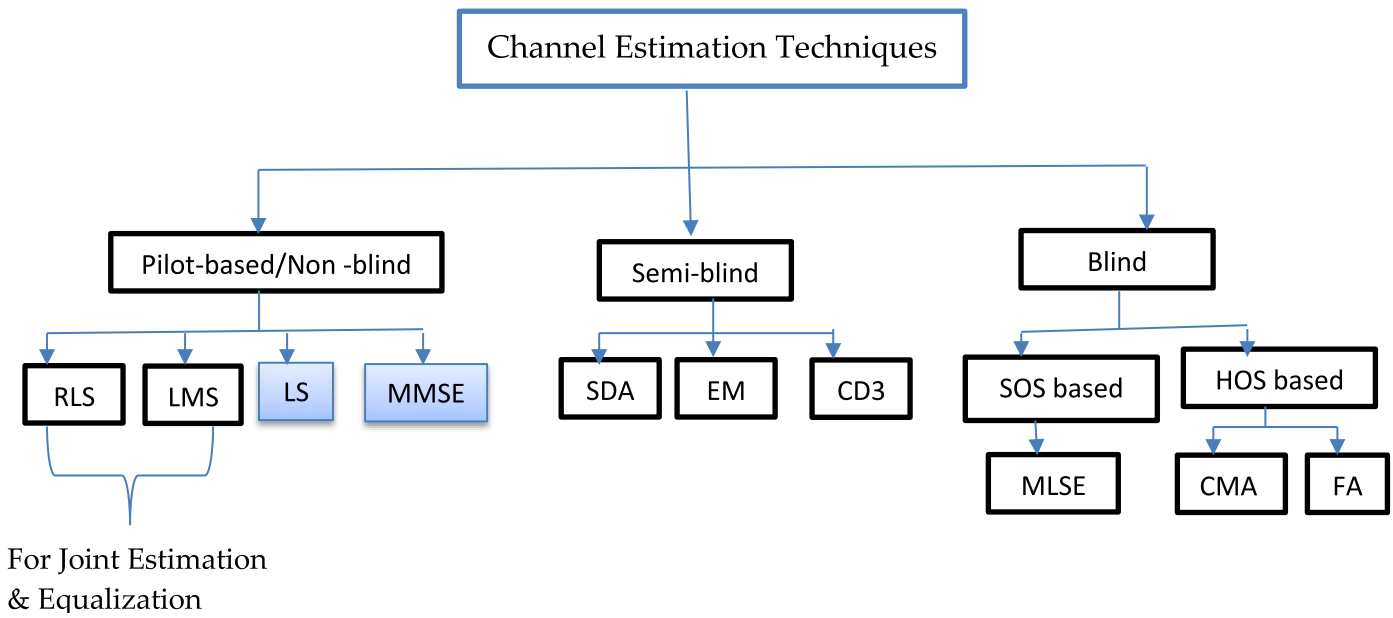


Figure 4.1 Classification of channel estimation techniques and algorithms

As shown in the figure above channel estimation techniques are divided in to three based on the requirement of training sequence. Non-blind techniques which needs training sequence for estimating of the channel, blind techniques that doesn't need any training sequence finally, and semi-blind channel estimation is the combination of pilot aided and blind channel estimation [46].

Semi blind/blind estimators and equalizers are believed to work unsatisfactorily in fading channels compared to training based methods, due to slow convergence or high computational complexity. In Ricean distribution (with Line of sight, LOS) environment semi blind/blind algorithms can be used. But in Rayleigh distribution channels it is better to use training based methods. Since Terrestrial transmission path is subject to numerous impacts such as echoes and multipath reception, AWGN and Doppler shift in case of mobile reception our main focus is on the pilot-based channel estimation techniques and algorithms. The brief presentation of blind and semi-blind techniques will be given in the last section of this chapter.

4.2.1 Pilot-based Channel Estimation Techniques

Pilot-based channel estimation is the method of doing channel estimation by using pilot sequence and symbol, which are inserted into some fixed positions of signals sent by transmitter [11]. Supervised (pilot-based) techniques simplify the receiver design, reduce the computational load and can be quite easily applied in wireless communications [5]. The pilot symbol sent by transmitter makes spectral efficiency and power utilization lower with the trade-off of quick response to the channel variation.

For pilot-based channel estimation of OFDM system, following three are required. Firstly, suitable pilot pattern needs to be considered. Secondly, pilot-based channel estimation algorithm with low complexity should be identified. Thirdly, proper demodulation method toward effective channel estimation has to be developed [11]. We will have brief presentation on pilot generation and pilot arrangement before discussing pilot-based, channel estimation algorithms.

4.2.2 Pilot Generation and Pilot Insertion

In order to obtain the channel information, pilot symbols are inserted in the information from transmitter, and the receiver get the channel information by using pilot symbols received. In essence, the problem of pilot pattern design is to determine where to insert the pilot and how closely between pilots. A suitable way of inserting could be calculated according to the known communication environment and estimated speed from the terminal [11].

Therefore, the most two important parameters of pilot, maximum Doppler shift f_m which determines the minimum coherent time, and maximum multipath time delay σ_{max} that decides the minimum coherent bandwidth [11]. The minimum limit of pilot symbols inserted is decided by Nyquist theorem. From Nyquist theorem, the interval of time domain S_t and frequency domain S_f should satisfy [11].

$$f_m.T.S_t.\leq \frac{1}{2}, \sigma_{max}.\Delta f.S_f.\leq \frac{1}{2} \tag{4.1}$$

Where Δf is banwidth of sub-carrier and T is period of the signal.

The pilot spacing is an important parameter generally this pilot are inserted in frequency domain and pilot spacing depends on the coherence frequency of the channel. Power allocation and modulation of the pilot data with regard to the factual data is an important parameter, if we increase pilot data power channel estimation accuracy increase [6].

4.2.3 Pilot Arrangement

Only restricted pilot data subcarriers are used for the primary channel estimation process. There are two ways to insert the pilots, these are in time and frequency domain [6].The estimation in this system is chosen by taking into account the required performance, computational complexity and time-variation of the channel [12].

Established along the pilot arrangement there are three forms of pilot arrangement considered: block type, comb type, and lattice type this three arrangements discussed in next sub sections the estimation accuracy can be amended by increasing the pilot density. They present some disadvantages. Main disadvantage it decreases the spectral efficiency [6].

4.2.3.1 Block Type Pilot Arrangement

As shown in Figure 4.2 pilots are placed for a block type of pilot arrangement. OFDM pilot symbols at all subcarriers are transmitted in a certain period for channel estimation [12].

A time domain interpolation using the pilots is performed to estimate the channel along the time axis. As explained in denoting S_t for the period of pilot symbols in time to combat time varying channel characteristics, the pilot symbols must be placed as frequently as the coherence time.

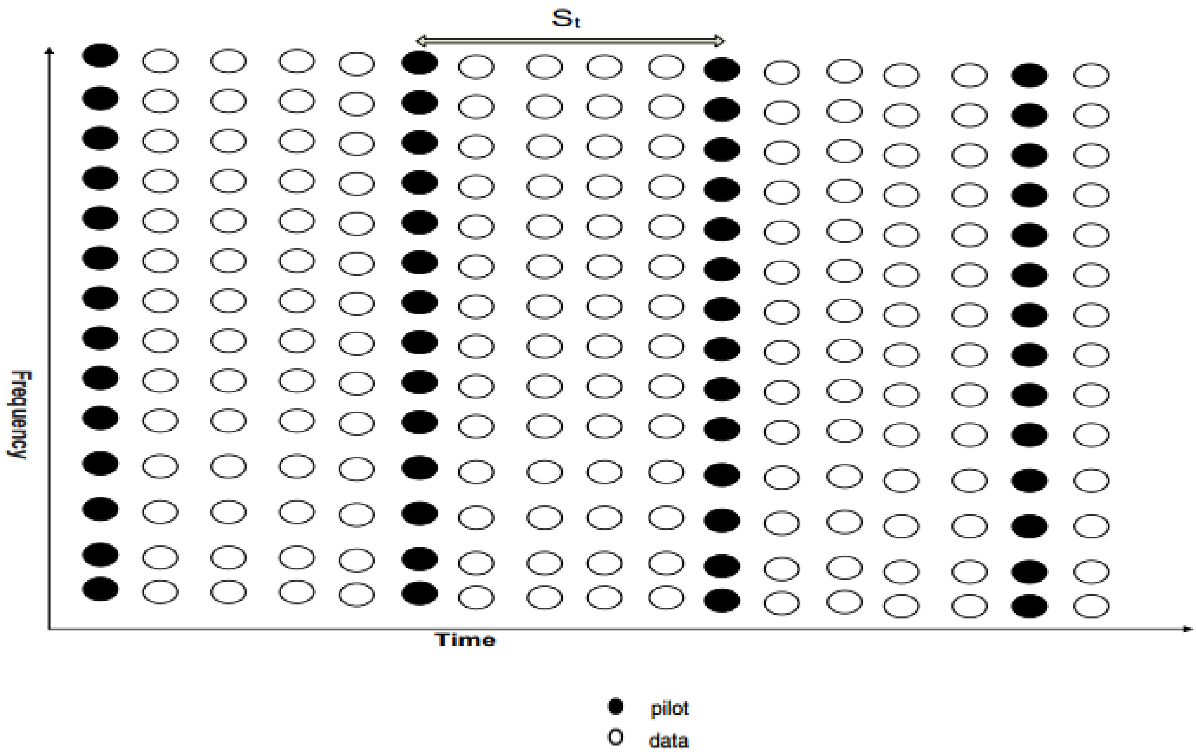


Figure 4.2 Block type pilot arrangement *[12, 13]*

The coherence time is given in an inverse form of the Doppler frequency $f_{Dopller}$ in the channel and the pilot symbol period must satisfy the following inequality:

$$S_t \leq \frac{1}{f_{Dopller}} \tag{4.2}$$

The block-type pilot arrangement is suitable for frequency-selective channels because pilot tones are inserted into all subcarriers of pilot symbols with a period in time. For the fast-fading channels, however, it might incur too much overhead to track the channel variation by reducing the pilot symbol period [12, 13].

4.2.3.2 Comp Type Pilot Arrangement

As shown in Figure 4.3 instead of all subcarrier the pilots are placed at periodically located subcarriers which are used to perform frequency domain interpolation to estimate the channel along the frequency axis.

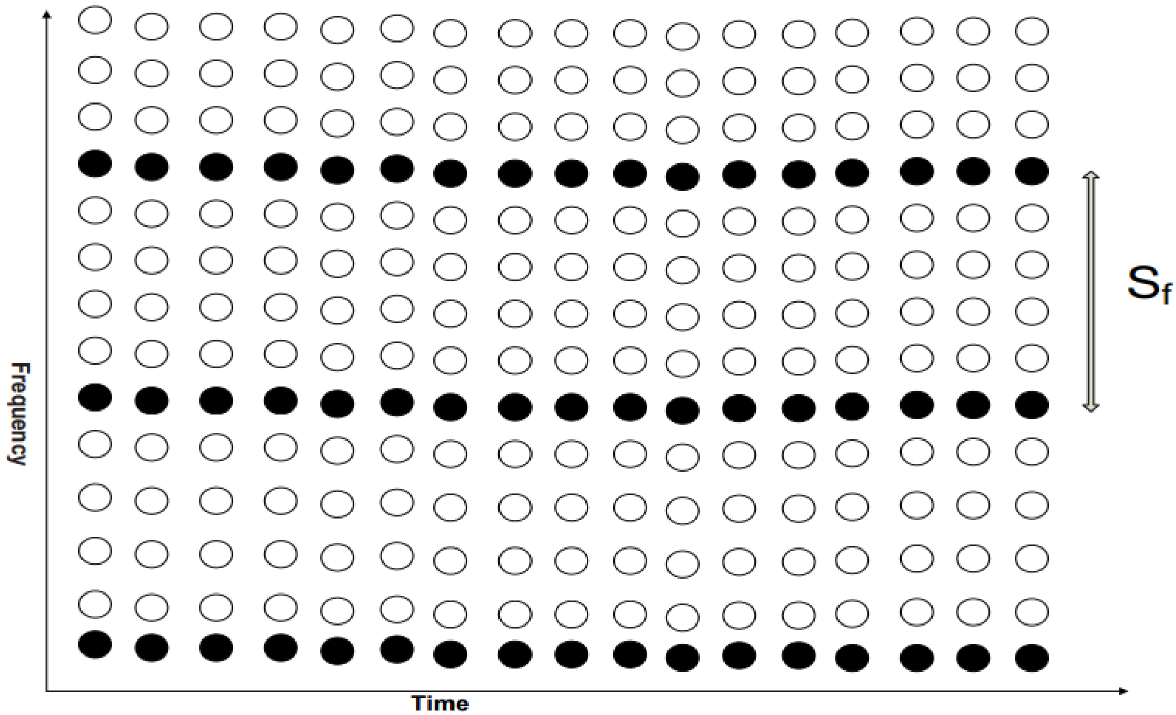


Figure 4.3 Comp type pilot arrangement*[12]*

Denoting S_f the period of pilot of tones in frequency, the pilot symbols must be placed as frequently as coherent bandwidths in order to track the frequency-selective channel characteristics. The coherence bandwidth is calculated by an inverse of maximum delay spread with the following inequality [13].

$$S_f \leq \frac{1}{\sigma_{max}} \tag{4.3}$$

In contrary to the block-type pilot arrangement, the comb-type pilot arrangement is suitable for fast-fading channels, but not for frequency-selective channels [47].

4.2.3.3 Lattice Type Pilot Arrangement

In this type pilot arrangement the pilots are inserted along both in time and frequency axes with certain frequency as shown in Figure 4.4. For channel estimation this arrangement can facilitate interpolation of both time and frequency domain.

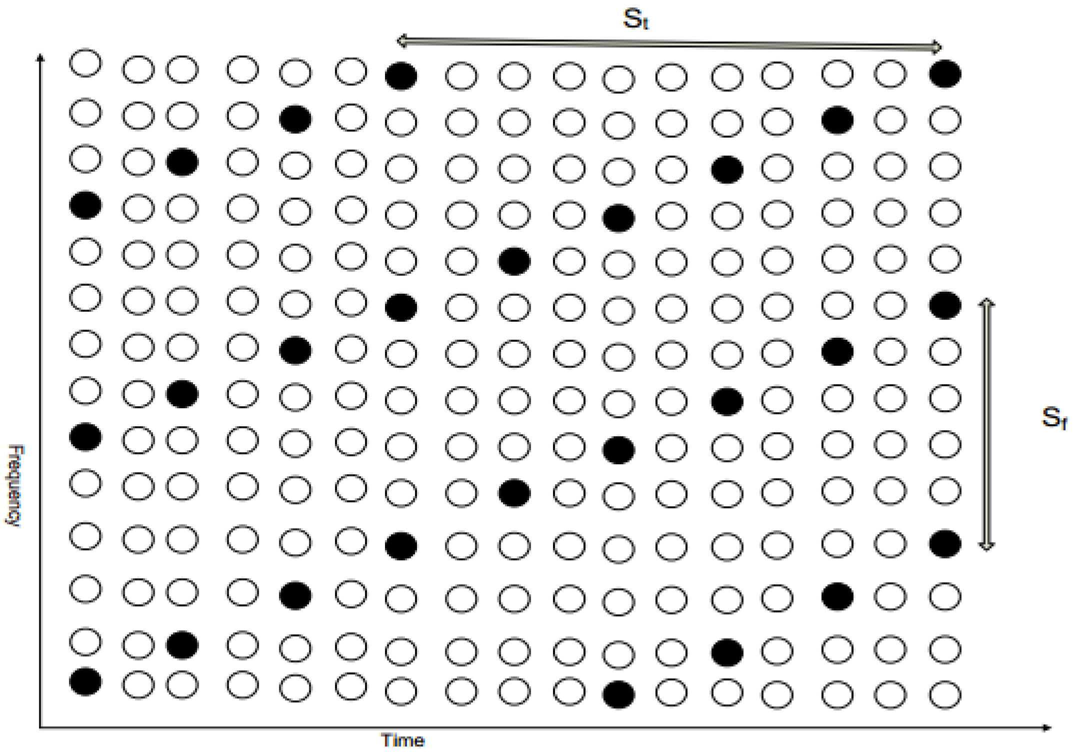


Figure 4.4 Lattice type pilot arrangement *[13]*

To keep track of both time-varying and frequency-selective channel characteristics, the pilot symbol arrangement must satisfy both inequalities in Equation 4.2 and 4.3

$$S_t \leq \frac{1}{f_{Doppler}} \text{ and } S_f \leq \frac{1}{\sigma_{max}} \quad (4.4)$$

Where $f_{Doppler}$ *and* σ_{max} Denote the Doppler spreading and maximum delay spread, respectively.

4.2.4 Pilot-based Channel Estimation Algorithms

As for channel estimation, there have been a variety of algorithms with different optimization criteria and levels of numerical complexity [3]. The most common pilot-based channel estimation algorithms are LMS, RLS, LS and MMSE. Iterative channel estimation algorithms LMS and RLS are suitable under fast fading channels [3].But these algorithms are commonly used for joint channel estimation and channel equalization.

When pilot symbols are used to estimate channel characteristics only, the LS and MMSE techniques are commonly used for channel estimation [12, 48, 49, 7, 8, 11].

Our assumption is all subcarriers are orthogonal. The representation of the diagonal matrix of training symbols for N subcarriers is as follows.

$$X = \begin{bmatrix} x(0) & 0 & 0 & 0 \\ 0 & x(1) & 0 & 0 \\ & & . & \\ & & . & \\ 0 & 0 & 0 & x(N-1) \end{bmatrix} \tag{4.5}$$

Where $X[k]$ denotes a pilot tone at the K^{th} subcarrier, with $E\{X[k]\} = 0$ and $Var\{X[k]\} = \sigma_x^2$, $k = 0, 1, 2, \dots, N-1$. By assumption all the subcarriers are orthogonal, X is a diagonal matrix. Representing the channel gain by $H[k]$ for each subcarrier k, the received training signal $Y[k]$ can be represented as

$$Y \equiv \begin{vmatrix} y(0) \\ y(1) \\ . \\ . \\ y(N) \end{vmatrix} = \begin{bmatrix} x(0) & 0 & 0 & 0 \\ 0 & x(1) & 0 & 0 \\ & & . & \\ & & . & \\ 0 & 0 & 0 & x(N) \end{bmatrix} \begin{bmatrix} H(0) \\ H(1) \\ . \\ . \\ H(N) \end{bmatrix} + \begin{bmatrix} Z(0) \\ Z(1) \\ . \\ . \\ Z(N) \end{bmatrix} \tag{4.6}$$

$$Y = XH + Z$$

Where H is the channel vector given as $H = [H(0), H(1), \dots, H(N-1)]^T$ and Z is noise vector given as $Z = [Z(0), Z(1), \dots, Z(N-1)]^T$ with $E\{Z(k)\} = 0$ and $var\{Z(k)\} = \sigma_Z^2$ $k = 0,1,, N-1$.

The main objective of this thesis is estimating the channel coefficient $\overline{H}$ using known pilot symbols. We will discuss on the most common pilot based channel estimation algorithms, LS, MMSE and LMS in the following sections.

4.2.4.1 Least Square Channel Estimation Algorithm

The goal of the channel least square estimator is to minimize the square distance between the received signal and the original signal.

For least square error estimation:

$$e = \left(Y_p - X_p H_p\right)^H \left(Y_p - X_p H_p\right)$$
$$e = Y_p^H Y_p - Y_p^H X_p H_p - X_p^H H_p^H Y_p + X_p^H H_p^H X_p H_p \tag{4.7}$$

Where

X_p is transmitted pilot signal vector.

Y_p is received pilot signal vector.

$(.)^H$ is conjugate transpose operation(Hermitian transpose of matrix).

H_p is channel coefficient at the pilot

The goal of the channel least square estimator is to minimize the estimation error.

$$H_{LS} = min(e) = min\left\{\left(Y_p - X_p H_p\right)^H \left(Y_p - X_p H_p\right)\right\} \tag{4.8}$$

For minimization of the error, e it is required to differentiate e with respect to H_p and evaluate with $\overline{H} = 0$ as follows:

$$\frac{\partial e}{\partial H_p} |_{\overline{H}} = 0$$

That is

$$-2Y_p^H X_p - 2\overline{H}^H X_p^H X_p = 0$$

$$Y_p^H X_p = \overline{H}^H X_p^H X_p \tag{4.9}$$

Multiplying both sides by $\left(X_p^H X_p\right)^{-1}$

$$\left(Y_p^H X_p\right)\left(X_p^H X_p\right)^{-1} = \overline{H}^H \left(X_p^H X_p\right)^{-1}$$

$$Y_p^H X X^{-1}\left(X_p^H\right)^{-1} = \overline{H}^H$$

$$\overline{H} = \left[\left(X_p^H\right)^{-1}\right]^H Y_p$$

$$\overline{H}_p = X_p^{\ -1} Y_p \tag{4.10}$$

Thus the LS estimation at pilot sub-carriers can be represented

$$\widetilde{H}_{LSp} = \frac{Y_p}{X_p}$$

$$\widetilde{H}_{LSp} = \left\{\widetilde{H}_p(0), \widetilde{H}_p(1) \ldots \widetilde{H}_p\left(N_{p-1}\right)\right\}^T = \left\{\frac{Y_p(0)}{X_p(0)}, \frac{Y_p(1)}{X_p(1)} \ldots \frac{Y_p\left(N_{p-1}\right)}{X_p\left(N_{p-1}\right)}\right\}^T \tag{4.11}$$

After having $\widetilde{H}_{LS}$ the LS channel estimate, we can calculate the MSE of the LS channel estimation algorithm by

$$\begin{aligned} MSE_{LS} &= E\left\{\left(H - \widetilde{H}_{LS}\right)^H \left(H - \widetilde{H}_{LS}\right)\right\} \\ &= E\{(H - X^{-1}Y)^H (H - X^{-1}Y)\} \\ &= E\{(X^{-1}Z)^H (X^{-1}Z)\} \\ &= E\{Z^H (XX^H)^{-1} Z\} \end{aligned} \tag{4.12}$$

$$MSE_{LS} = \frac{\sigma_Z^2}{\sigma_X^2}$$

Where σ_Z^2 and σ_X^2 are the noise variance and pilot signal variance respectively. As we can observe from Equation 4.10 above the reduction of σ_Z^2 the MSE performance will be consistently improved. But it may suffer noise enhancement, that is, the signal power σ_X^2 is too small in the deep faded channels. Hence, the MSE performance of LS channel estimation will be significantly degraded. Although LS technique has the above problem, as a result of mathematical simplicity it has been widely used for channel estimation [15].

The LS channel estimation at the data is obtained after channel interpolation is performed. Different channel estimation techniques will be presented in section.

The merit of LS estimator is its low complexity while it ignores the impact of the noise and it doesn't exploit neither the time nor the frequency correlation of the channel. Thus, it is more sensitive to the interference of the noise.

4.2.4.2 MMSE Based Channel Estimation

The MMSE estimator employs the second-order statistics of the channel conditions to minimize the MSE. As we can observe from Figure 4. 5 the error of channel estimation is e

$$e = H - \overline{H}$$

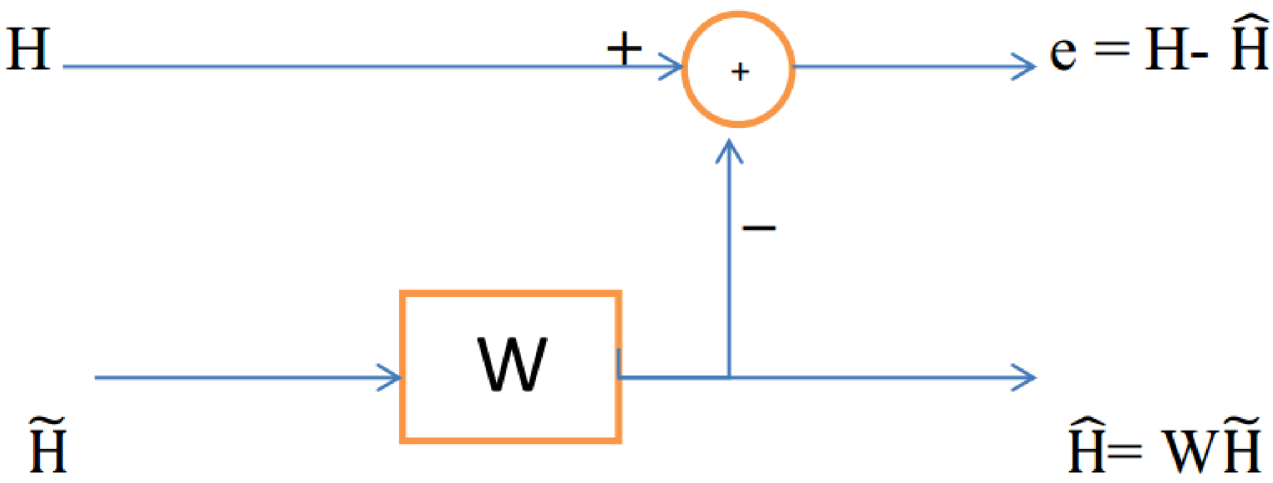


Figure 4.5 Minimum Mean Square Error channel Estimation *[7, 12]*

Where H is actual channel estimation and $\overline{H}$ is a raw channel estimation and W is weight matrix. The MSE of channel estimation.

$$E\{|e|^2\}= E\{|H - \overline{H}|^2\}=E\{(H - \overline{H})(H - \overline{H}^H)\} \tag{4.13}$$

Where $E\{\}$ *is* the expectation. Since the channel and AWGN are not correlated, we can rewrite Equation 4.13 as

$$\overline{H}_{MMSE=}R_{HY}R_{YY}^{-1}Y \tag{4.14}$$

Let us denote the auto-covariance matrixes of H,Y by R_{HH} and R_{YY} respectively, across covariance matrix between H and Y by R_{HY}. Let σ_N^2 is the noise-variance, since channel and AWGN are not correlated, we could get

$$R_{HY} = E\{HY^H\}$$
$$=E\{H(HX+W)^H\}$$
$$=\{HH^HX^H + HW^H\}$$
$$=\{HH^HX^H + 0\}$$
$$R_{HY} \quad =R_{HH}X^H \tag{4.15}$$
$$R_{YY} = E\{YY^H\}$$
$$=E\{(HX+W)(HX+W)^H\}$$
$$=E\{HXH^HX^H + HXW + WH^HX^H + WH^H\}$$
$$=E\{HH^H\}XX^H + 0 + 0 + E\{WW^H\}$$
$$R_{YY} \ =XR_{HH}X^H+\sigma_N^2 I_N \tag{4.16}$$

If R_{HH} and σ_N^2 are known to the receiver, channel impulse response could be calculated by MMSE estimator as below

$$\overline{H}_{MMSE=}R_{HY}R_{YY}^{-1}Y$$
$$=R_{HH}X^H\left(XR_{HH}X^H + \sigma_N^2 I_N\right)^{-1}X\overline{H}_{LS}$$
$$=R_{HH}\left(R_{HH} + \sigma_N^2\left(X^HX\right)^{-1}\right)]^{-1}\overline{H}_{LS} \tag{4.17}$$

The performance of MMSE estimator is much better than LS estimator, especially under the lower SNR. However, because of the required matrix inversion the computation is very complex when the number of subcarriers of OFDM system increases. Therefore, an important drawback of MMSE estimator can be the high computational complexity [11].

In order to eliminate the interference of the noise and increase the accuracy of the estimator, we should use MMSE criterion to design the channel estimation algorithm. It is so complex that it is nearly impractical to implement. The simplified linear minimum mean square error criteria for channel estimation and is formulated as

$$\widetilde{H}_{LMMSE} = R_{HH}\left(R_{HH} + \frac{\beta}{SNR}I\right)^{-1}\widetilde{H}_{LS}$$

Where $R_{HH} = E\{HH^T\}$ $\beta = E\{|X_k|^2\}\{|1/X_k|^2\}$ and SNR= $E\{|X_k|^2\}/\delta_n{}^2$

R_{HH} *is the autocovariance matrix of the channel* $\delta_n{}^2$ *denotes the noise variance* $E\{|n_k|^2\}$, β *is constant* decided by constellation diagram and I is the identity matrix. In a practical system, the channel autocorrelation and SNR can be set to known factors at the receiver in advance. The channel autocorrelation calculation and SNR estimation affect the performance.

Compared with the LS method, MMSE method this is better but more complex because this must estimate the inverse of the channel estimation matrix therefore more calculations. Additionally, MMSE algorithm requires knowledge of channel covariance and noise variance, which are assumed to be known as a priori knowledge. The LS algorithm does not need any knowledge about the channel autocorrelation [50].

4.2.4.3 LMS Based Channel Estimation

The LMS algorithm was devised by Widrow and Hoff in 1959 in their study of a pattern-recognition machine known as the adaptive linear element, commonly referred to as the Adaline. The LMS algorithm is a stochastic gradient algorithm in that it iterates each tap weight of the transversal filter in the direction of the instantaneous gradient of the squared error signal with respect to the tap weight in question [51].

The LMS is a search algorithm in which a simplification of the gradient vector computation is made possible by appropriately modifying the objective function. The LMS algorithm, as well as others related to it, is widely used in various applications of adaptive filtering due to its computational simplicity.

The block diagram of adaptive LMS algorithm is given below.

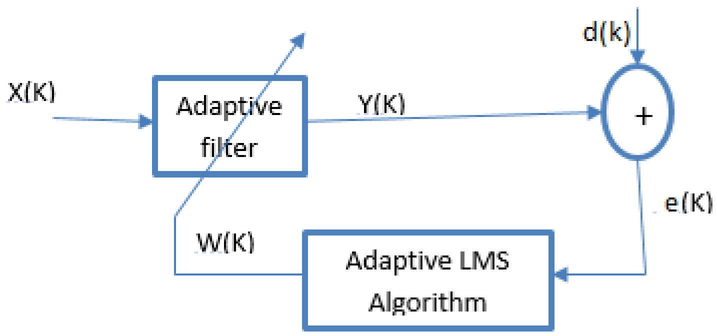


Figure 4.6 LMS algorithm presentation *[2]*

Where $X\ (k)$ is the input signal, $Y\ (k)$ is the filter output, $d\ (k)$ is the desired input signal and $e(k)$ is the error between $d(k)$ and $y(k)$.

Adaptation process involves the following three steps

1. Compute output of the filter

$$y\ (k)=\sum_{i=0}^{N-1} W_{i(k)}\ x\ (k-i) = \boldsymbol{w}^T(k)\ x\ (k) \qquad (4.18)$$

2. The value of the error estimation is calculated using $d\ (k) - y\ (k)$
3. The tap weights of the FIR vector are updated in preparation for the next iteration by

$$w(k+1)= w(k)+2\mu e(k)\ x(k) \qquad (4.19)$$

This method tries to adjust the filter parameter in such a way that minimizes the MSE between the output of the filter and the desired signal. The optimum weight coefficients are updated adaptively depending on the error.

$$\begin{bmatrix} New\ coefficent \\ vector \end{bmatrix} = \begin{bmatrix} old\ coefficient \\ vector \end{bmatrix} + \begin{bmatrix} Adaptation\ gain \\ vector \end{bmatrix} . \begin{pmatrix} error \\ signal \end{pmatrix} \qquad (4.20)$$

The convergence characteristics of the LMS algorithm are examined in order to establish a range for the convergence factor that will guarantee stability. The convergence speed of the LMS is shown to be dependent on the Eigen value spread of the input signal correlation matrix [52].

$$0 < \mu < \frac{1}{\lambda_{max}} \tag{4.21}$$

Where, λ_{max} is the largest eigenvalue of input signal autocorrelation.

The choice of μ as explained above ensures that the mean value of the coefficient vector approaches the optimum coefficient vector w_0. It should be mentioned that if the matrix **R** has a large Eigen value spread, it is advisable to choose a value for μ much smaller than the upper bound. As a result, the convergence speed of the coefficients will be primarily dependent on the value of the smallest Eigen value, responsible for the slowest mode in equation above.

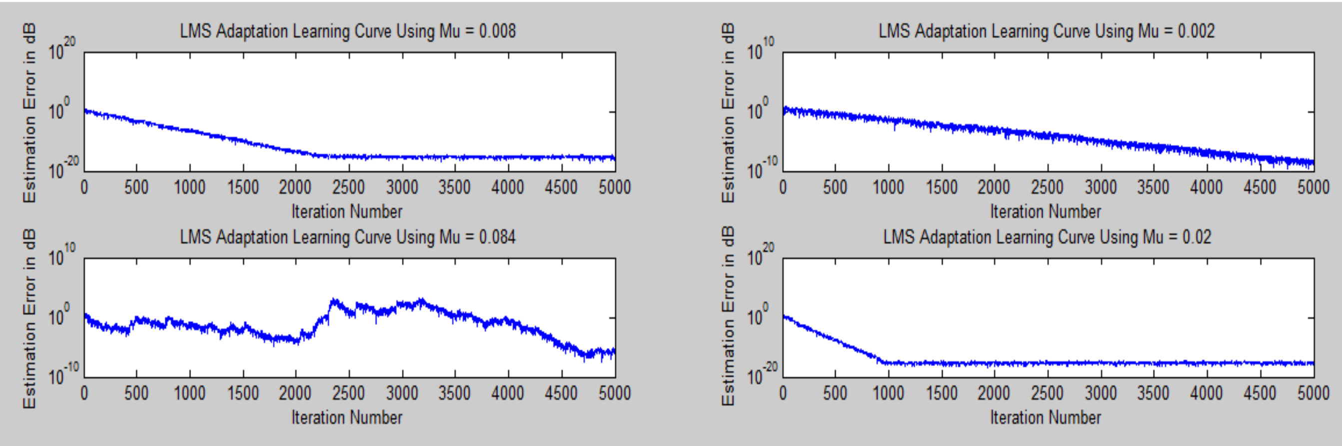


Figure 4.7 Convergence and stability of LMS for different μ

As we can see from Figure 4.7 the step size value should select carefully as stability and speed of convergence are dependent on that. The algorithm converges after 1000 iteration when we take step size of 0.02 and converges at 5000 iteration in step size of 00.002 but when we increase the step size to 0.084 it becomes unstable.

Draw Backs of LMS Algorithm

The convergence to the optimum solution and speed at which this convergence takes place are greatly dependent on the value of the step-size parameter. Also the step-size parameter affects the stability of algorithm. The step-size parameter should be in the range.

$0<\mu<\frac{1}{\lambda max}$ where λ_{max} is the maximum of the Eigen values of autocorrelation matrix R. Higher value provides faster convergence but if it exceeds certain bound then the algorithm will be unstable. However its convergence speed is highly dependent on the Eigen value spread of the input covariance matrix. For highly correlated inputs the LMS algorithm has a slow convergence which requires long training sequences and therefore low transmission speeds. Another drawback of the LMS is the trade-off between convergence speed and steady-state error since both are controlled by the same parameter, the step-size.

In order to eliminate these drawbacks, the class of Variable Step-Size LMS (VSSLMS) algorithms was introduced. There is Complementary Pair Variable Step-Size LMS algorithms highly increase the speed of convergence while reducing the trade-off between the convergence speed and the output error [53]. The VSS-LMS algorithm involves one additional step size update equation compared with the standard LMS algorithm.

4.2.4.4 RLS Based Channel Estimation

RLS adaptive filter is an algorithm which recursively finds the filter coefficients that minimize a weighted linear least squares cost function relating to the inputs signal. This is in contrast to other algorithm such as the LMS algorithm that aim to reduce mean square error [44].The RLS algorithm for a pth order RLS filter can be summarized as,

Main parameters

ρ =Filter order

λ =Forgetting factor

δ =Vvalue of initialized $P(0)$

Initialization $w_n = 0$ and $P(0) = \delta^{-1}I$ where I is the $(P+1)$ by $(P+1)$ identity matrix

The weight oupdate can be given by the following Equation

$$w(n) = w(n-1)+\alpha(n)g(n) \quad ,\text{n=0,1,2,...,} \tag{4.22}$$

Where

$\alpha(n) = d(n) - w(n-1)^T x(n)$

$g(n) = p(n-1)z(n)\{\lambda + x^T(n)p(n-1)x(n)\}^{-1}$

RLS algorithm is the the fastest But it requires more complicated calculation.

Table 4.1 comparison of different pilot based channel estimation algorithms

Algorithms	Advantages	Dis advantages	comment
LMS	Easier to implement than RLS	Requires training sequence.	Commonly used for joint channel estimation and equalization.
RLS	Always converges faster than LMS	Requires training sequence, computationally complex	Commonly used for joint channel estimation and equalization.
LS	Very easy to implement	Doesn't correlate with noise	Can be used for explicit channel estimation
MMSE	Fast convergence than LS	Very high complexity i.e. Requires matrix inversion.	Can be used for explicit channel estimation.

Since the objective of this thesis is for extracting the channel coefficients independently the simulation concentrates on LS and MMSE channel algorithms only.

4.2.5 Channel Interpolation Techniques

The channel estimation based on comb type pilot insertion, an interpolation technique is necessary in order to estimate channel at data sub-carriers by using the channel information at pilot sub-carriers [46].

Pilot data symbols are inserted at a regularly Interval in actual block of data. In specified location here the channel estimation is done by using the interpolating the pilot data to the nearest pilot data channel estimation is done by using the pilot interpolated data [6].

According to the pilot allocation schemes discussed in 4.2.3, the pilots will be inserted in the time or frequency domain to estimate the channels for a particular time instant or subcarriers. However, the channels for data symbols are unknown to the receiver except the pilot channel estimates [54]. The interpolation techniques are needed to estimate the channels between subcarriers or time slots.

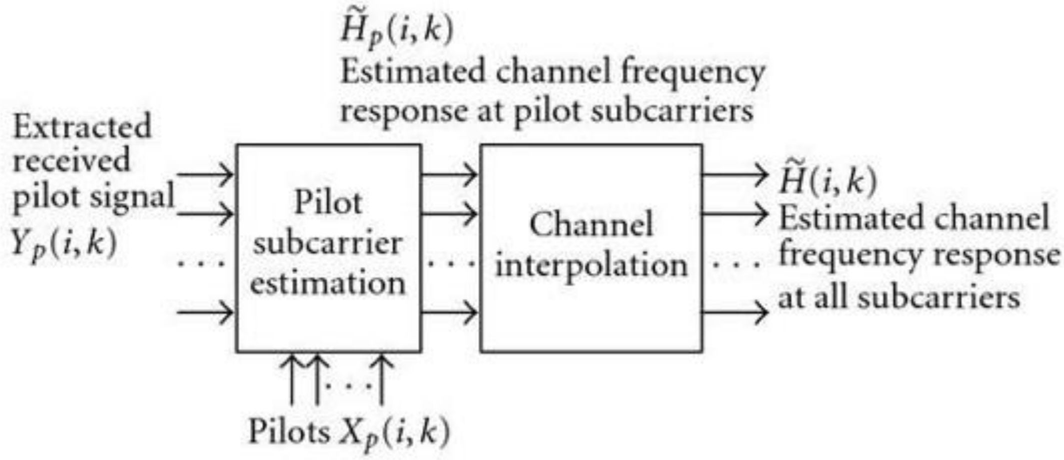


Figure 4.8 channel estimation based on comp-type pilots *[8]*

As we can in the Figure 4.8 above the channel coefficients at the pilots $\overline{H}_p$ are estimated from the transmitted and received pilots followed by channel interpolation to estimate the coefficients at the transmitted data $\overline{H}_{d.}$

In this paper, channel estimation is performed in two steps: The first step is to estimate the channel frequency coefficients at the pilot symbols positions using LS and MMSE estimator.

The second step using the estimates of the channel frequency coefficients we then interpolate over channel frequency coefficients corresponding to the data symbols [55].

The interpolation can be linear, second-order, low-pass, and spline cubic and time-domain interpolation. Second-order interpolation has been shown to perform better than linear interpolation. Low-pass interpolation is the one with the best performance among all where all the mentioned techniques are compared with each other [5].

4.2.5.1 Piecewise Constant Interpolation

We call it piecewise constant interpolation (or nearest neighbor interpolation). It allocates same pilot values to near data subcarriers [14].

4.2.5.2 Linear Interpolation

Linear interpolation is a simple method to outperform piecewise constant interpolation which estimates the channels between two pilots with the aid of linear approximation. The optimal number of pilots for a given BER can be determined in advance for OFDM systems with numerical evaluation as in [54]. The channel estimation at the data subcarrier is obtained by estimation of response of two adjacent pilot sub-channels. But the precondition is linearity of transmitted functions of adjacent sub-channels. The linear interpolation is shown as below,

$$\overline{H}(\mathrm{k})= \overline{H}(\mathrm{m}N_f + l)=\left(\overline{H}_p(m+1) - \overline{H}_p(m)\right)\frac{1}{N_f}+\overline{H}_p(m),\mathrm{m}=0,1\ldots N_p - 1 \tag{4.23}$$

In Equation 4.8 above, in the linear interpolation estimation only first two nearest data points are interpolated this pilot interpolated data points used estimate the channel this very simple method [6].

4.2.5.3 Second-order Interpolation

Here we used three pilot carriers to estimate the second order Interpolation. For second-order interpolation, the MSE performance will be better than that of the linear interpolation with complexity increase, and the idea behind it is similar to the linear interpolation except that it employs a second-order approximation [54].

Second order interpolation is better than linear interpolation, where the channel estimation at the data subcarrier is calculated by used linear combination of three adjacent pilots [11]. Figure 4.7 shows the diagram of second order interpolation.

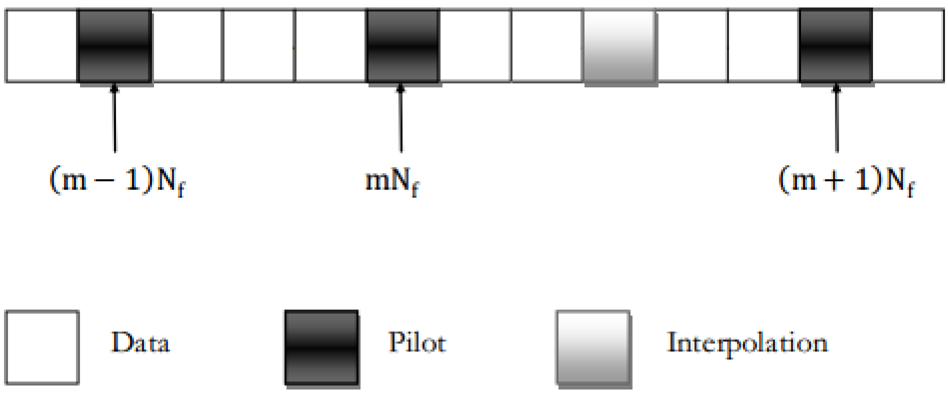


Figure 4.9 Sketch map of second order interpolation *[11]*

Theoretically, high order interpolation yields better channel estimation because of using more pilots. And the channel estimation will be close to the true channel response. But computation complexity also is increased as increasing of order. The channel estimation of second order interpolation is given by

$$\bar{H}(\mathrm{k}) = \bar{H}(\mathrm{m}N_f + l) = c_1\bar{H}_p(m-1) + c_0\bar{H}_p(m) + c_{-1}\bar{H}_p(m1), \mathrm{m}=0,1\ldots N_p - 1 \quad (4.24)$$

The coefficients are defined as

$$\text{Where} \begin{cases} c_1 = \frac{\alpha(\alpha-1)}{2} \\ c_0 = -(\alpha-1)(\alpha+1), \alpha = \frac{l}{N_f} \\ c_{-1} = \frac{\alpha(\alpha+1)}{2} \end{cases}$$

4.2.5.4 Cubic Spline Interpolation

The cubic spline interpolation method produces a smooth and continuous polynomial fitted to given data points [11, 6]. The fundamental idea behind spline cubic interpolation is based on draw smooth curves through a number of points [55], which is given by:

$$\bar{H}(\mathrm{k})=\bar{H}(\mathrm{m}N_f+l)=\alpha_1\bar{H}_p(m+1)+\alpha_0\bar{H}_p(m)+N_f\alpha_1\bar{H}'_p(m+1)-N_f\alpha_0\bar{H}'_p(m),\mathrm{m}=0,1\ldots N_p-1 \quad (4.25)$$

Where $\bar{H}'_p(m)$ is the first derivative of $\bar{H}_p(m)$ and

$$\alpha_1=\frac{3(N_f-l)^2}{{N_f}^2}-\frac{2(N_f-l)^3}{{N_f}^3}$$

$$\alpha_1=\frac{3l^2}{{N_f}^2}-\frac{2l^2}{{N_f}^3}$$

Although cubic spline interpolation with higher order interpolation can be used for better interpolation accuracy, the performance improvement is not obviously proven [11].

4.2.5.5 DFT-Based Interpolation

The DFT-based interpolation technique is the output of the Fourier transform of the channel impulse response. The DFT-based interpolation effectively removes the effects of noise outside the maximum channel delay spread or the length of multipath channel. The implementation of the DFT-based interpolation is also straightforward compared to the linear or second-order interpolation, but the length of multipath channel must be known to the receiver [54]. DFT based interpolation technique only requires FFT and IFFT computation as well as the length of multipath, so it is a very common approach used in channel estimation for OFDM systems [54].

4.2.5.6 MMSE Interpolation

Compared to other interpolation techniques, MMSE interpolation may be the most effective way to estimate the channels between pilots with the aid of statistical information on the channels such as the channel correlation and SNR, as well as additional computation of the matrix inversion. However, it can achieve an excellent performance as compared to linear, second-order and DFT-based interpolation [54]

4.2.5.7 Low-Pass Interpolation

The low-pass interpolation method is performed by inserting zeros into the original sequence and then applying a low-pass finite-length impulse response (FIR) filter, which allows the original data to pass through it without any changing. This method also interpolates such that the mean-square error between the interpolated points and their ideal values is minimized [55].

Numerical Example on Interpolation and channel estimation

To have better understanding of channel estimation and channel interpolation lets have one numerical example by considering the following system parameters.

Parameters	Values
N_{FFT}	16
N_p	4
Pilot position	[0 4 7 11]
Transmitted pilots$[X_p(0)\ X_p(1)\ X_p(2) X_p(3)]$	[1 1 1 1]
Received pilots $[Y(0)\ Y_p(1)\ Y_p(2) Y_p(3)]$	[0.2 2.5 2.8 1.2]

From the above parameters we can find channel estimation values at data subcarriers using piecewise constant interpolation, linear interpolation, and second-order polynomial interpolation.

First we have to find the channel frequency response at the pilot position using LS estimator. From Equation 4.11, LS channel estimation determining the channel frequency response is calculated at pilot positions as follows:

H_{LSp}=$\{H(0), H_p(1) \dots H_p(N_{p-1})\}^T$=$\left\{\frac{Y_p(0)}{X_p(0)}, \frac{Y_p(1)}{X_p(1)} \dots \frac{Y_p(N_{p-1})}{X_p(N_{p-1})}\right\}^T$=$[0\ .2.\ 2.5\ 2\ .8.\ 1.\ 2]$ and plot the channel coefficients as shown in Figure 4.10.

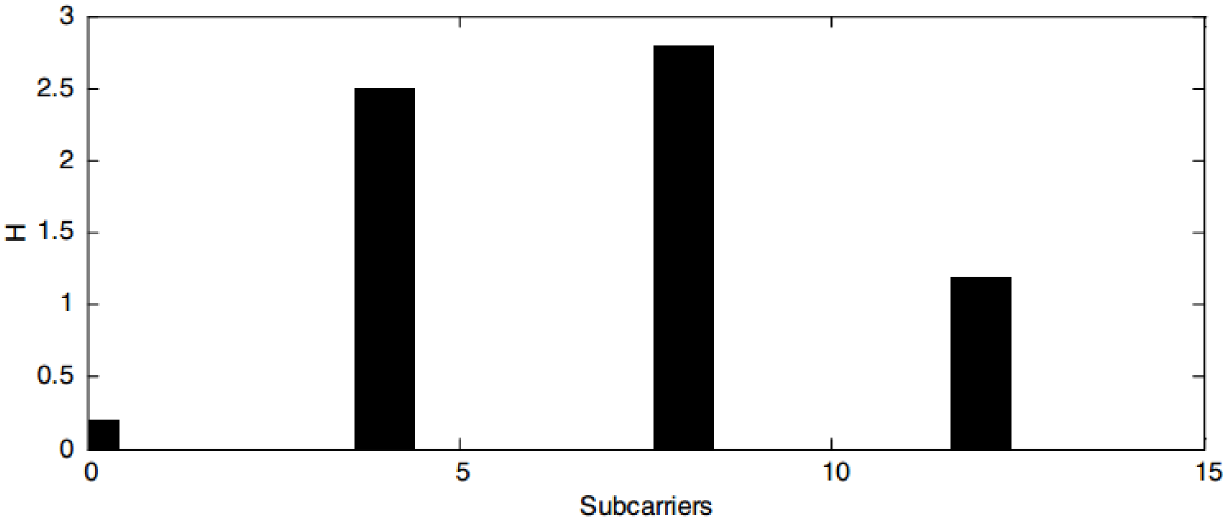


Figure 4.10 Channel response at pilot position

The piecewise constant interpolation method allocates same pilot values to near data Subcarriers. Thus, we can easily obtain channel estimation values as shown in Figure 4.11.

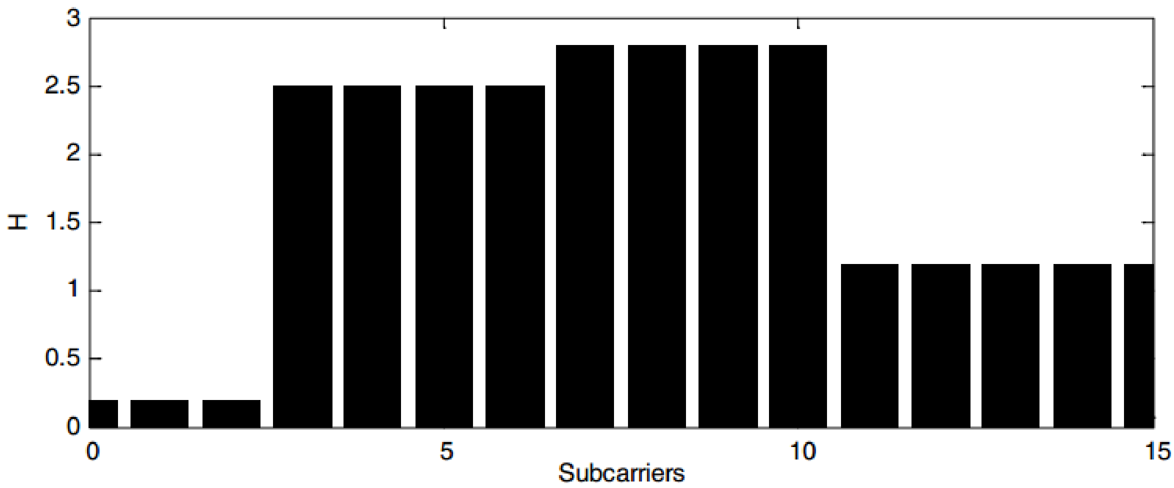


Figure 4.11 Piecewise constant interpolation

The linear interpolation method allocates channel estimation values on the straight line between two pilots. From (4.23) the channel estimation value $\bar{H}$(k) between $H_p(0), H_p(1)$ are calculated as follows:

$$D = N_{FFT/N_p} = \frac{16}{4} = 4$$

$$mD \leq (m+1)D \; when \; m = 0 \rightarrow 0 \leq k < 4$$

$$0 \leq d < D \rightarrow 0 \leq d < 4$$

$$\bar{H}(\mathrm{k})=\frac{d}{4}\Big(H_p(1)-H_p(0)\Big)+H_p(0)$$

$$[\bar{H}(0)\ \bar{H}(1)\ \bar{H}(2)\ \bar{H}(3)\ ]=[0.2\ 0.78\ 1.35\ 1.92]$$

In the same way, the remaining channel estimation values $\bar{H}(\mathrm{k})$ are calculated as follows:

$mD \leq (m+1)D\ when\ m=1 \rightarrow 4 \leq k < 8$

$$\bar{H}(\mathrm{k})=\frac{d}{4}\Big(H_p(2)-H_p(1)\Big)+H_p(1)$$

$$[\bar{H}(4)\ \bar{H}(5)\ \bar{H}(6)\ \bar{H}(7)\ ]=[2.5\ 2.58\ 2.65\ 2.73]$$

$mD \leq (m+1)D\ when\ m=2 \rightarrow 8 \leq k < 12$

$$\bar{H}(\mathrm{k})=\frac{d}{4}\Big(H_p(3)-H_p(2)\Big)+H_p(2)$$

$$[\bar{H}(8)\ \bar{H}(9)\ \bar{H}(10)\ \bar{H}(11)\ ]=[2.8\ 2.4\ 2\ 1.6]$$

Figure 4.12 illustrates linear interpolation.

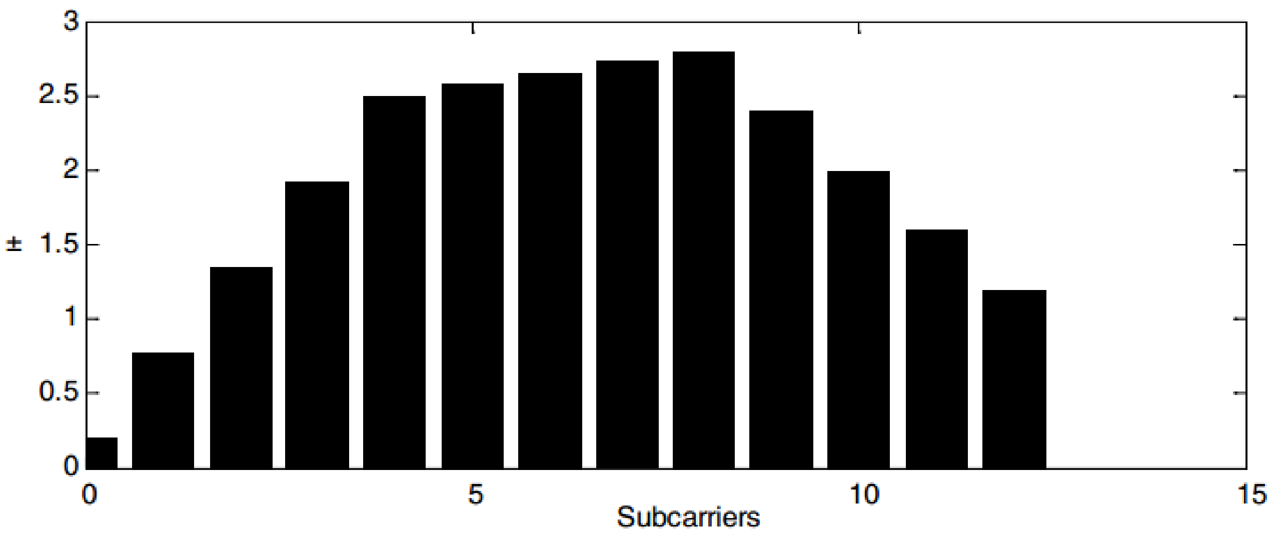


Figure 4.12 Linear interpolation

The second-order polynomial interpolation method allocates the channel estimation values using linear combination of the weighted pilot values. From (4.23) the channel estimation value

$$mD \leq (m+1)D\ when\ m=1 \rightarrow 4 \leq k < 8$$

$$0 \leq d < D \rightarrow 0 \leq d < 4$$

$$\alpha = \frac{d}{16}$$

$c_1 = \frac{\alpha(\alpha-1)}{1}$, $c_0 = -(\alpha - 1)(\alpha + 1), c_{-1} = \frac{\alpha(\alpha+1)}{2}$

$$\bar{H}(\text{k}) = C_1 H_p(0) + C_0 H_p(1) + C_{-1} H_p(2)$$

$$[\bar{H}(4)\ \bar{H}(5)\ \bar{H}(6)\ \bar{H}(7)\ ] = [2.5\ 2.58\ 2.65\ 2.71]$$

In the same may, the remaining channel estimation values $\bar{H}$(k) are calculated as follows:

$mD \leq (m + 1)D\ when\ m = 2 \rightarrow 8 \leq k < 12$

$$\bar{H}(\text{k}) = C_1 H_p(0) + C_0 H_p(1) + C_{-1} H_p(2)$$

$$[\bar{H}(8)\ \bar{H}(9)\ \bar{H}(10)\ \bar{H}(11)\ ] = [2.8\ 2.76\ 2.7\ 2.64]$$

Figure 4.13 illustrates the result of the second-order polynomial interpolation method. In the linear interpolation method and the second-order polynomial interpolation method, several interpolants were not estimated because the pilots are not enough to cover all subcarriers.

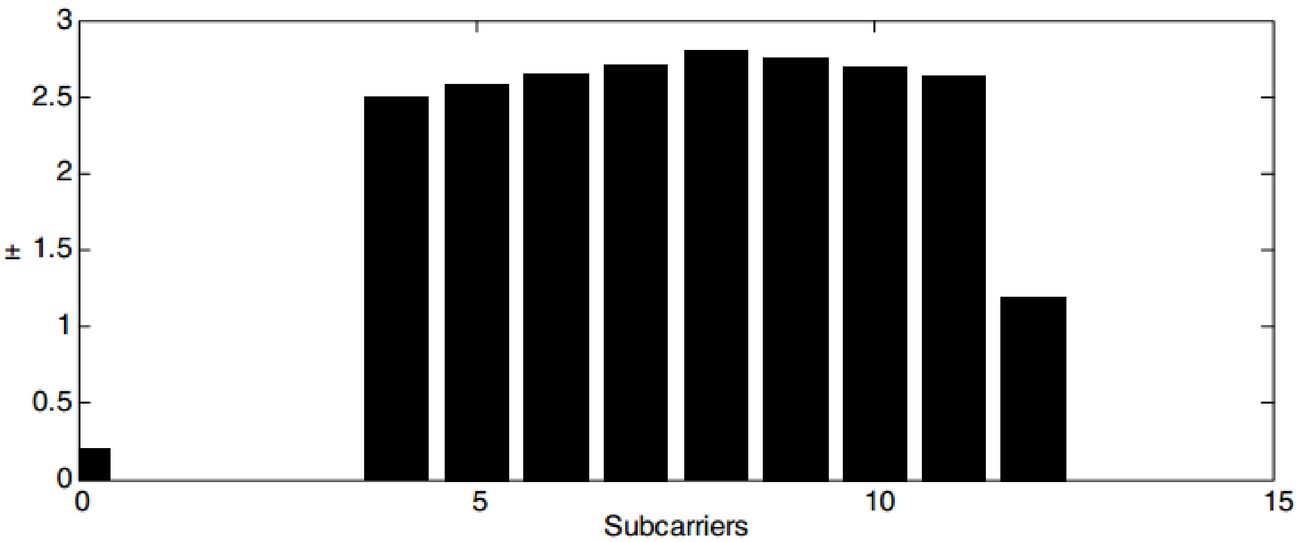


Figure 4.13 Second-order polynomial interpolation

4.2.6 Blind Channel Estimation Techniques

Blind channel estimation is focusing on the correlation between the data sent and received, without knowing the information of the transmitted data. Although it yields higher spectral and power efficiencies by using blind channel estimation, it needs require longer data records and entail higher complexity. Hence it is suitable for slow varying channel [11].

The BCE methods can be classified into HOS -based techniques and SOS-based techniques [7]. This techniques typically employ FA, CMA and MLSE estimation algorithms. But, they usually suffer from complexity and convergence problems. This thesis is concentrated on pilot based channel estimation techniques only.

4.2.7 Semi-Blind Channel Estimation Techniques

Semi-Blind technique combines both blind and non-blind (with initial training-based channel estimation and next channel tracking). Assumes an intermediate position and relies partly on training and partly on the use of channel statistics. Compared to purely training-based schemes, the number of training symbols can be saved a lot, while the channel estimation quality is still conserved [56]. This technique require less computational complexity than blind methods and fewer training symbols than training-based methods, making them attractive for practical implementation.

Semi-blind algorithms can improve the performance of blind algorithms by exploiting the knowledge of both known symbols and properties of the transmitted signals. The objective of semi-blind channel estimation algorithms is to get better performance than blind algorithms while requiring fewer known symbols than training based channel estimation algorithms [50].

SDA, EM, and CD3 algorithms are the suitable approaches to implement this technique. However, the performance is not comparable to that of pilot assisted channel estimation as the Doppler frequency, and there by, the speed of the receiver increases [5].We do not detail further on this method because this thesis concentrates on the pilot based methods.

4.2.8 Selection Criteria of Channel Estimation Algorithms

For one estimation technique, there may exist several adaptive algorithms that could be used to adjust the weight vector. The choice of one algorithm over another is determined by various factors which include:

- **Rate of convergence**: - This is defined as the number of iterations required for the algorithm, in response to stationary input, to converge to the optimum solution. A fast rate of convergence allows the algorithm to adapt rapidly to a stationary environment of unknown statistics. Furthermore, it enables the algorithm to track statistical variations when operating in a nonstationary environment [2].
- **Computational complexity**:- refers to the number of operations (i. e., multiplications, divisions, and additions/subtractions) required to update the filter from one time instant to the next.
- **Numerical properties**:-The numerical accuracy and stability of the algorithm refers to error minimization capability of the algorithm. When an algorithm is implemented numerically, inaccuracies are produced due to round-off' noise and representation errors in the computer. These kinds of errors influence the stability of the algorithm [2].
- **Misadjustment:-** For an algorithm of interest, this parameter provides a quantitative measure of the amount by which the final value of the MSE, averaged over an ensemble of adaptive filters, deviates from the optimal minimum MSE [2].

4.2.9 Complexity Analysis of Channel Estimators

The computational complexity is one of the important factors of the channel estimator algorithm performance. Before determining the complexity of the channel estimation algorithms, a number of general rules will be introduced, namely, the complexity of a matrix multiplication, matrix addition and matrix inversion. Table 4.1 shows the general overview of matrix calculation rules to determine the computational complexity. Where capital O notation represents the computational complexity and n represents the matrix size.

Table 4.2 matrix calculation rules overview *[57, 48]*

Operation	Input	Output	Algorithm	Complexity
Matrix multiplication	Two $n \times n$ matrices	One $n \times n$ matrices	**Schoolbook matrix multiplication**	$\boldsymbol{O(n^3)}$
			Strassen algorithm	$O(n^{2.807})$
			Coppersmith-Winograd algorithm	$O(n^{2.376})$
Matrix multiplication	One $n \times m$ matrix and One m$\times$ p matrix	One $n \times p$ matrices	**Schoolbook matrix multiplication**	$\boldsymbol{O(nmp)}$
Matrix inversion	One $n \times n$ matrix	One $n \times n$ matrix	**Gauss-Jordan elimination**	$\boldsymbol{O(n^3)}$
			Strassen algorithm	$O(n^{2.807})$
			Coppersmith-Winograd algorithm	$O(n^{2.376})$
Discrete Fourier Transform (DFT)	One $n \times n$ matrix	one number with at most $O(n(\log n))$bits	**General DFT algorithm**	$\boldsymbol{O(n(\log n))}$

N.B the bolded algorithms are used in this thesis for complexity analysis.

4.2.9.1 Computational Complexity for MMSE Estimator

The computational complexity for MMSE algorithm depends on Equations (4.13) - (4.16) defining the transfer $\boldsymbol{h_{MMSE}}$ function impulse response, cross co-relation of g and$y$,$\boldsymbol{R_{Hy}}$.

Mathematical calculation of complexity for $\boldsymbol{R_{Hy}}$:

$$O(\boldsymbol{R_{Hy}}) = O(\boldsymbol{R_{HH}}F^H X^H) = n^3 + n^2 = n^2(n+1) \tag{4.26}$$

Where $\boldsymbol{R_{HH}}$ is a $n \times n$ matrix, F is a $n \times n$ matrix, x is a $1 \times n$ matrix. Number of operation needed for square matrix $(n \times n) * (n \times n)$ is $O(n^3)$and matrix multiplication of $(n \times n) * (n \times 1)$ is $O(n * n * 1)$ or $O(n^2)$

So the computational complexity for $\boldsymbol{R_{Hy}}$ becomes $n^2(n+1)$

Mathematical calculation of the complexity of $\boldsymbol{R_{yy}}$

The computational complexity for $\boldsymbol{R_{yy}}$ is

$$O(\boldsymbol{R_{yy}}) = O(xF\boldsymbol{R_{HH}}F^H x^H + \delta_n^2 I_n) = n^2 + n^3 + n^3 + n^2 + n^2 = n^2(2n+3) \tag{4.27}$$

Number of operation needed for square matrix $(n \times n) * (n \times n)$ is $O(n^3)$ and matrix multiplication of $(n \times n) * (n \times 1)$ is $O(n * n * 1)$ or (n^2)

Mathematical calculation of complexity of$\boldsymbol{\overline{H}_{MMSE}}$:

$$\boldsymbol{O(\overline{H}_{MMSE}) = O(R_{Hy}R_{yy}^{-1}y) = n^3 + n^2 + n^3 = n^2(2n+1)} \tag{4.28}$$

Number of operation needed for square matrix $(n \times n) * (n \times n)$ is $\boldsymbol{O(n^3)}$and matrix multiplication of $(n \times n) * (n \times 1)$ is$O(n * n * 1)$ or $O(n^2)$ and for inverse operation $\boldsymbol{n^3}$.

Computational complexity for LS estimator

From Equation (4.10) and Table 4.2 we get the complexity of LS estimator as:

$$\mathrm{O(H_{LS}) = O(X^{-1}Y) = n + n^2 = n(n+1)} \tag{4.29}$$

Here one inverse operation and one multiplication operation. So for inverse operation we need to consider n operation because matrix size of is n x1 and for multiplication need n^2 operation.

Analysis summary

Table 4.3 shows the complexity analysis of all estimators-'. After analysis the complexity evaluation it can be concluded that LS estimator gives the lowest complexity because it consists of only one multiplication and one inverse operation and the MMSE estimator gives the highest computational complexity.

Table 4.3 Computational complexity analysis *[58]*

Estimation scheme	Number of operations needed	Complexity	Comments
LS estimator	$n(n+1)$	Lowest	Simple matrix multiplication and inverse operation
MMSE estimator	$n^2(2n+3)$	Very High	Large number of matrix multiplications
LMS	$2n+1$	LOW	Doesn't require matrix inversion
RLS	$4n^2$	High	More complex than LMS but easier than MMSE

Chapter Five

Analysis and Simulation

In the previous Chapters the theoretical background and basic principles of channel estimation techniques and algorithms is developed. In this chapter will present MATLAB simulation results and discussions on the obtained results. MATLAB-based simulation is used to investigate the performances of the LS and MMSE pilot-based channel estimation algorithms. In this thesis we extract the channel coefficients using LS and MMSE algorithms in frequency selective channels and adjust the shifted phases using LMS algorithm in flat fading channels.

5.1 Simulation Model and Parameters

The whole DVB-T2 system model is very complex. In the aim of this analysis we can treat with simplified simulation model with reduced complexity. The simplified simulation model for this thesis is given in Figure 5.1 below. As we can see from the below block diagram the whole system includes transmitter, channel and receiver blocks.

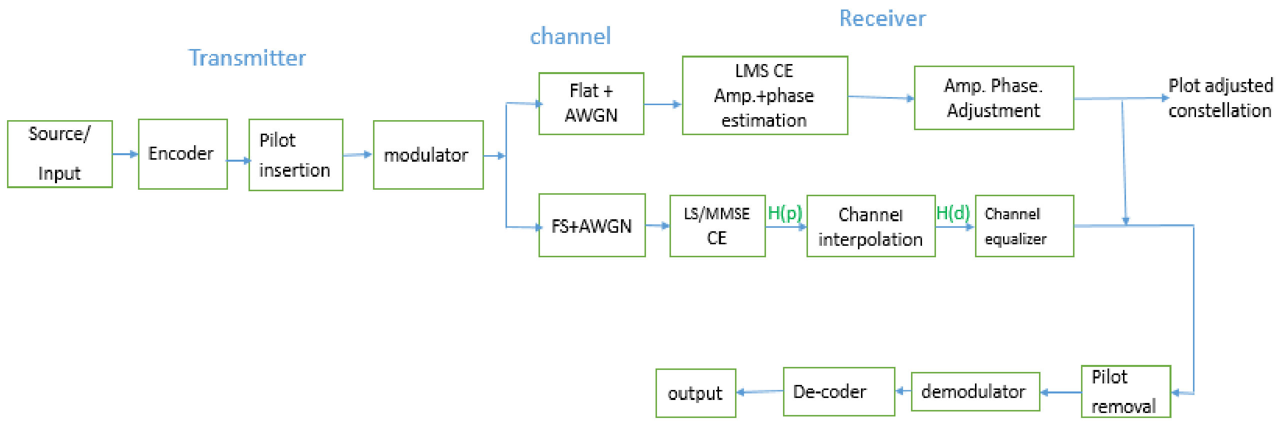


Figure 5.1 Simulation Block diagram

The transmitter bock includes source encoder, pilot generator and modulator, the input is encoded before pilot insertion and modulation. Each complex data-valued complex symbol is then assigned to a data subcarrier location. Also reference complex symbols (or pilots) are generated and inserted at the reserved pilot subcarrier location for channel estimation and channel interpolation purpose. In this thesis block type pilot arrangement is used and tested in ¼, 1/8 and 1/16 the data of pilot ratio.

Source: - Two data types are used for the simulation namely a randomly generate bits and an image files. Before encoding and modulation the image file is converted to data using image to data converter MATLAB function.

The Channel:-The medium from transmitter to receiver is expressed as multipath fading unknown channel. In this we consider frequency selective and flat fading channels further corrupted by an AWGN.

At the receiver side, mainly the reverse operation of transmitter blocks is done. The received signal is estimated to compensate the distorted amplitude and shifted phase using LMS algorithm in flat fading channel since ISI is absent and estimated weights are given to channel equalization in frequency selective channel.

Channel estimation:- in frequency selective channel using LS and MMSE algorithms channel is estimated at the pilot subcarrier locations and is followed by interpolation, to estimate channel at data subcarrier locations. But in flat fading channel the shifted phase can be estimated and adjusted since ISI is absent.

Channel interpolation:-after channel estimation the channel coefficients are interpolated at the data using interpolation techniques. In this thesis the cubic spline interpolation techniques is selected based on BER vs SNR performance and complexity. The estimated coefficients $H(d)$ are used by channel equalization for further channel compensation and ISI removal.

The parameters used for the simulation are summarized in Table 5.1 below. The tradeoff on the choice of the parameters for DVB-T2 standard are presented in Chapter three of section 3.6.8. Besides the requirement of DVB-T2 standard before extraction of the channel coefficient using LS and MMSE channel estimation algorithm different performance comparison are done in order to select best channel interpolation techniques, modulation technique, pilot size and appropriate channel length. From the different performance comparison the optimum values are selected for the implementation of channel estimation as listed in the table below.

Table 5.1 Simulation parameters

Parameters	Value used
Modulation/Demodulation	PSK
Modulation order	4
Estimation algorithm	LS and MMSE
IFFT/FFT size	1k
Number of carriers	256
Pilot ratio	¼, 1/8 and 1/16 of the data
Pilot arrangement	Block type
SNR	15dB
Channel model	TU6
Number of channel taps	6,8
Cyclic prefix length	8,4
Channel interpolation	cubic spline
Channel BW	8MHZ
Sampling Fre.	10000Hz
LMS step size	0.01
Source image size	120 × 120 pixels

5.2 Simulation Results and Analysis

The MATLAB simulation results for frequency selective and flat fading channels using the above listed parameters are given in the following section in the form of figures and tables. The main purpose of channel estimation in frequency selective channel is to extract the channel coefficients for further channel equalization.

5.2.1 Performance Comparison of Different Channel Models

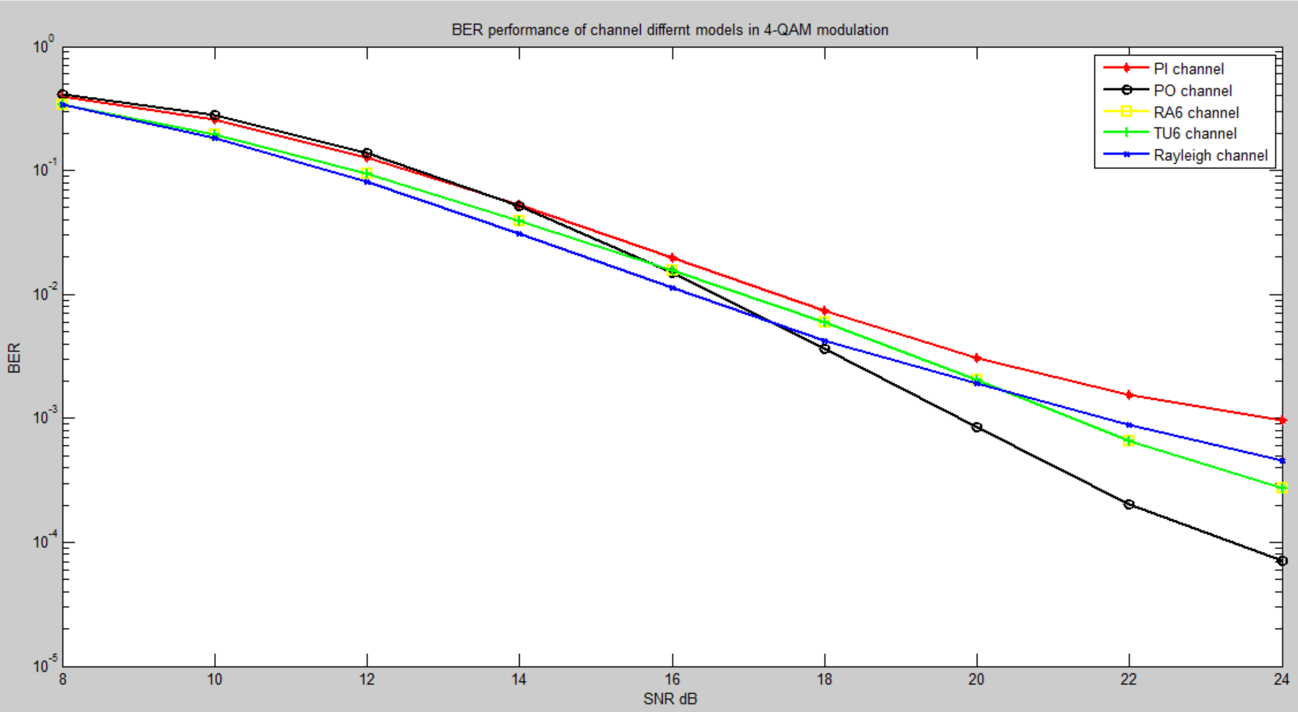


Figure 5.2 BER performance of channel models in 4-QAM

As it shown in Figure 5.2 above PI channel which has the worst performance between all 5 channels and PO has best performance at SNR of greater than 14dB. The main difference between PI and PO is channel models is in the lengths of the impulse responses and the delays of output paths. In both channels, the influence of attenuation is significant and the PI has worse results in simulation because in broadcasting the indoor channel model is affected by both indoor and outdoor attenuations. Channel models TU6 and RA6 has almost the same performance at all SNR levels. In this thesis for channel coefficient extraction TU6 is selected due to performance and suitability for DVB-T2 standards.

5.2.2 Performance Comparison of Different Channel Interpolation Techniques

In pilot based channel estimation after estimating the channel the weights should be interpolated using the channel interpolation techniques. The performance comparison of different channel interpolation techniques is given in the Figure 5.3. The interpolation is performed based on the channel estimates obtained by LS or MMSE channel estimation as previously described in Chapter four.

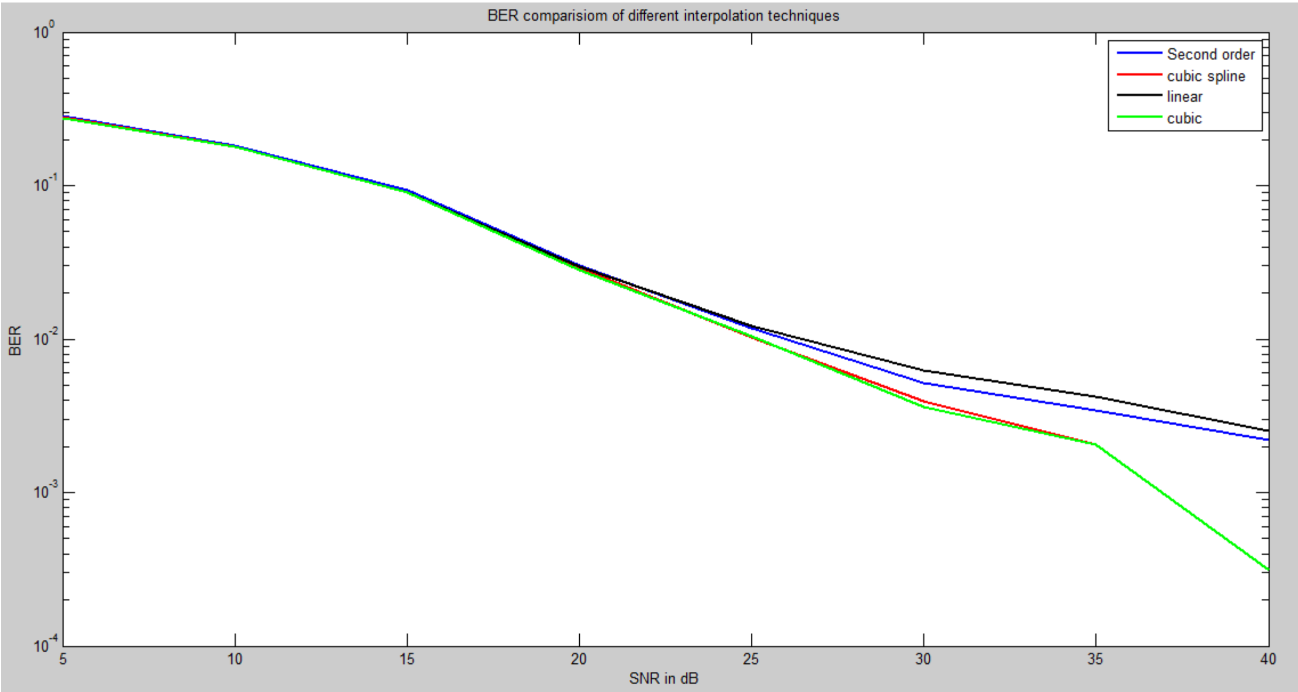


Figure 5.3 BER comparison of different channel interpolation techniques

As it is shown in Figure 5.3 above, as the order of interpolation technique increases the performance becomes better but complexity increases. The less complexity linear interpolation technique has good performance for SNR value less 20dB but for high SNR the performance of Linear interpolation is significantly degraded whereas the performance of cubic and cubic spline interpolations perform well at high SNR value. Depending on their performance and computational complexity in this thesis work is simulated using the cubic spline interpolation for further implementation of channel estimation using LS and MMSE estimation algorithms.

5.2.3 *BER vs SNR Comparison of LS and MMSE Channel Estimation Algorithms*

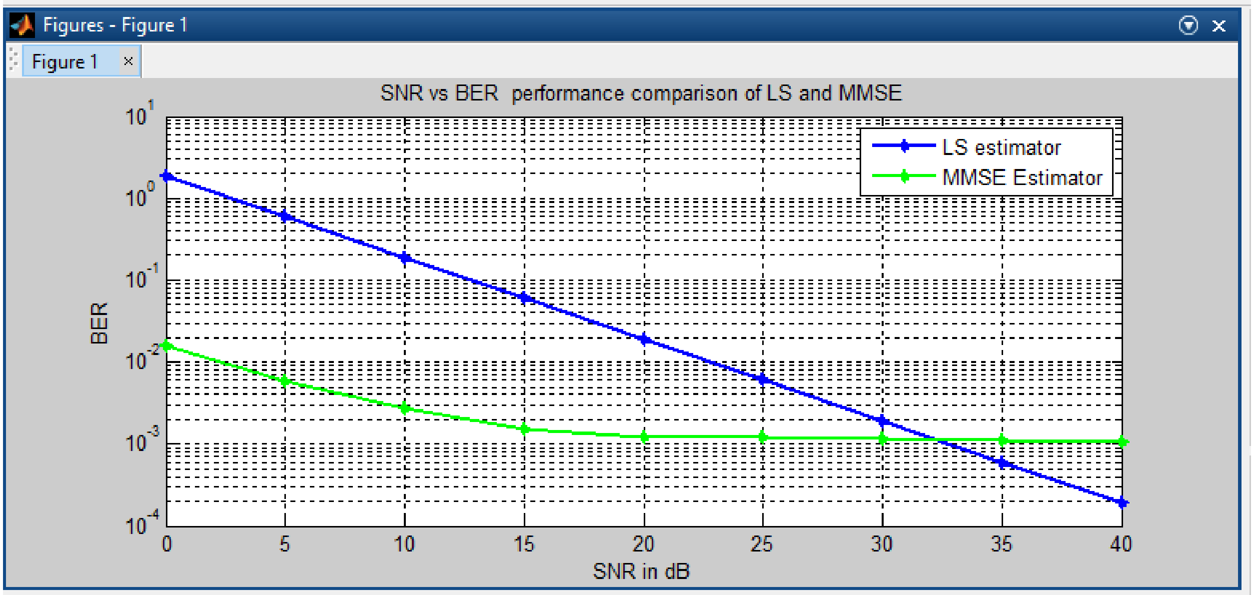


Figure 5.4 BER vs SNR plot of LS and MMSE channel estimators L=6 and CP=4

The performance of the two types of estimators LS and MMSE based on BER vs SNR is given in the above Figure 5.4 is simulated at the length of the channel (L) 6 and cyclic prefix (CP) length of 4. The simulation result shows that MMSE performs better than LS channel estimation algorithm at lower SNR but for higher SNR values MMSE loses its performance and LS estimator seems to perform better than MMSE for this range of SNR values. When the channel length is longer than the CP length MMSE provides better performance only for low SNR values and begins to lose its performance for higher SNR values. The only drawback of MMSE at lower SNR is its computational complexity. On the other hand LS is very simple to implement than MMSE but worst in performance.

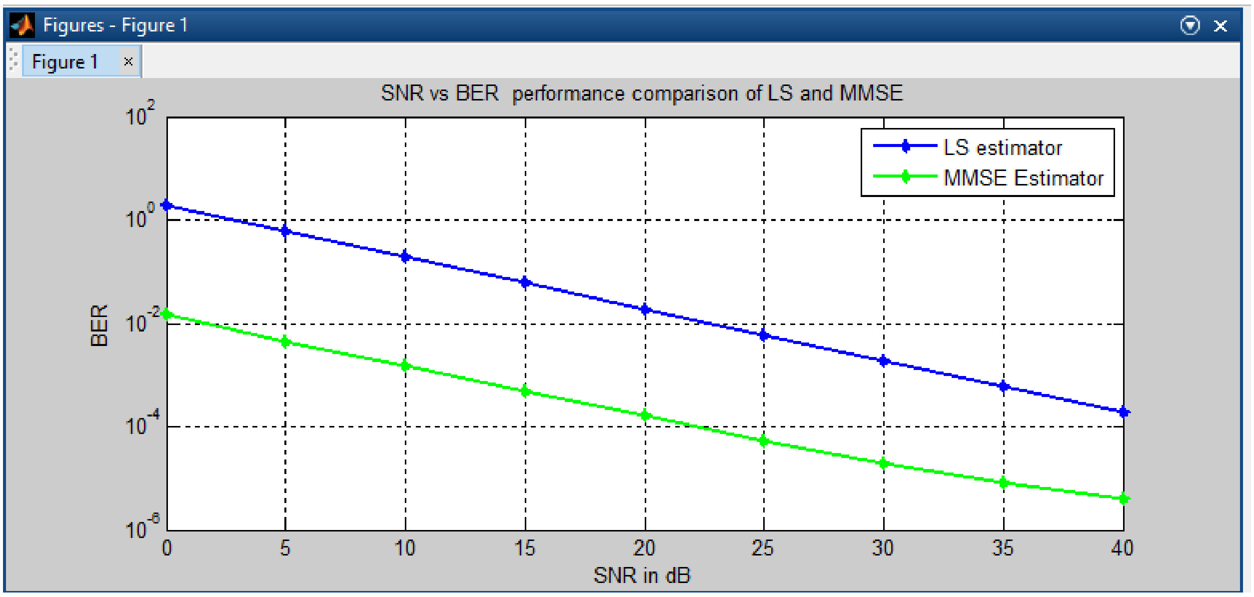


Figure 5.5 BER vs SNR plot of LS and MMSE channel estimators L=6 and CP=6

In this case, the cyclic prefix is equal to the channel MMSE estimation technique is better than the LS estimator but its complexity is higher due to the channel correlation and the matrix inversion lemma.

5.2.4 Performance Analysis of MMSE and LS Channel Estimation in QAM and PSK

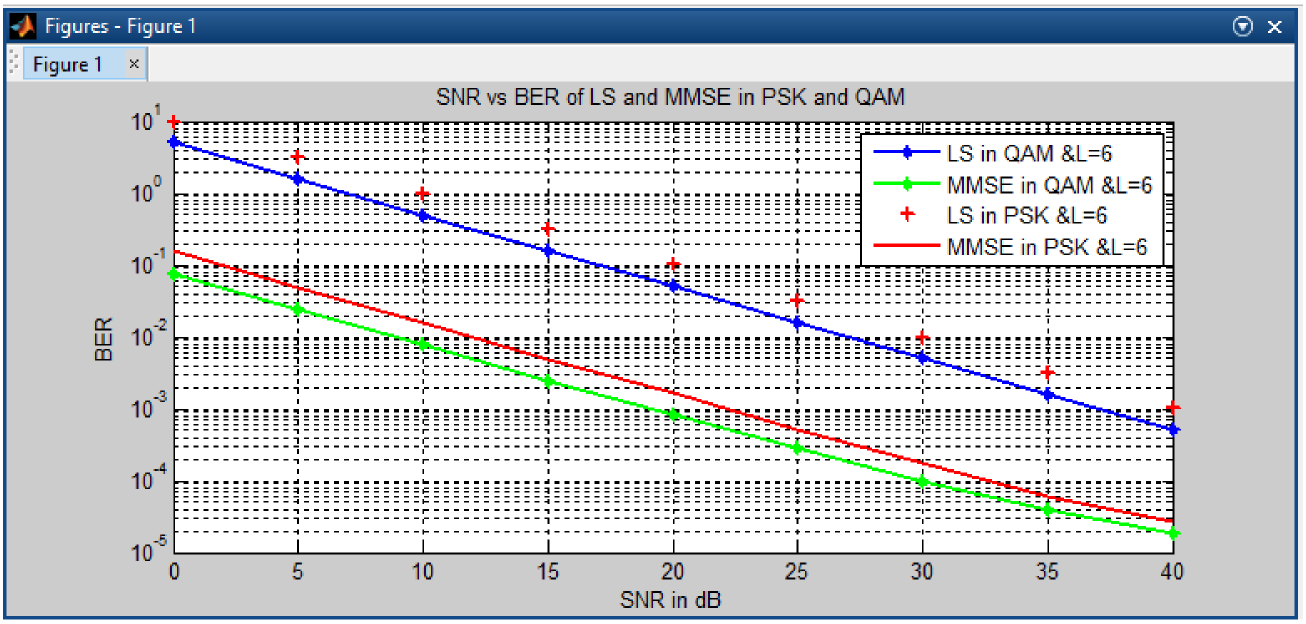


Figure 5.6 LS and MMSE comparison in PSK and QAM modulations

Figure 5.6 shows that the performance of LS and MMSE Estimators in PSK is better than QAM modulation technique. Although QAM is widely used in wireless communication system in terms of data carrying capacity but QAM is more susceptible to noise compared to QPSK. Even though QAM is seen to have the same spectrum and bandwidth efficiency as QPSK but still it is difficult demodulate in the presence of noise. The performance of MMSE is better than LS in both modulation techniques. In DVB-T2 standard required modulation schemes are QPSK, 16QAM, 64QAM and 256QAM, so in this thesis QPSK modulation technique is used in frequency selective channels to extract the channel coefficients and in flat fading channels for the shifted phase adjustment.

5.2.5 Performance Analysis of MMSE and LS Channel Estimation in Different Pilot Size

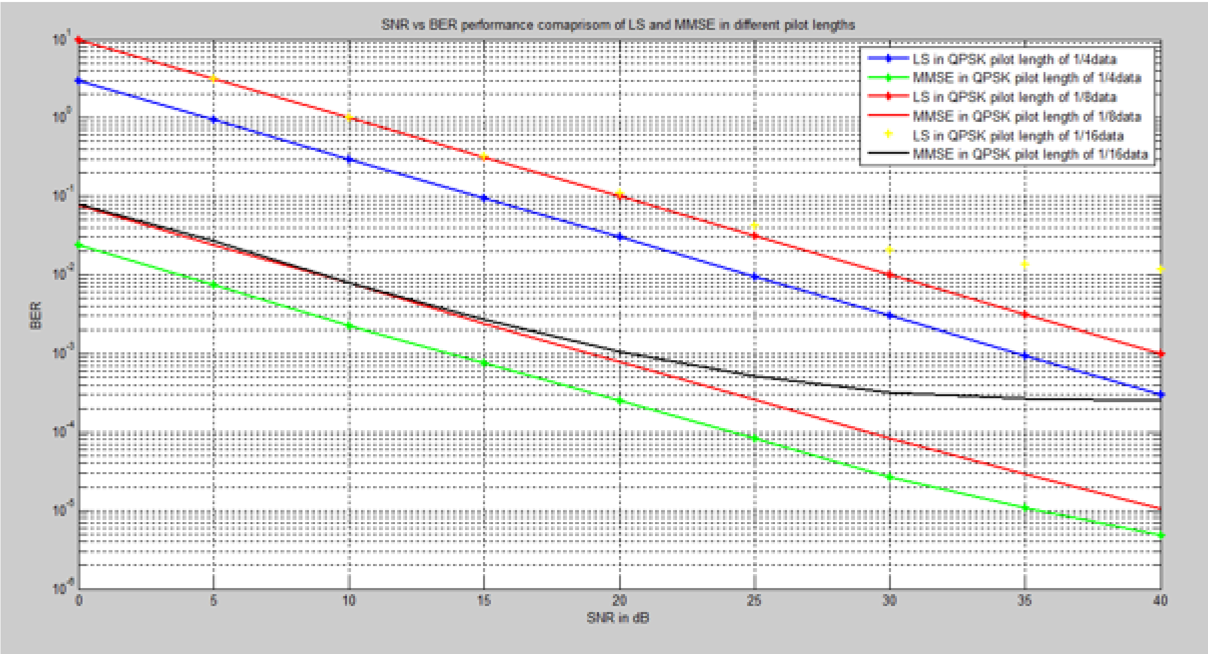


Figure 5.7 Performance comparison of LS and MMSE for different pilot sizes

The performance of both channel estimation algorithms is different in different pilot sizes i.e. becomes better when pilot size increases. As we can see from Figure 5.7 above using MMSE algorithm we can achieve BER of 10^{-3} at 13dB when we use pilot size ¼ of the data but when we use 1/8 of the data the SNR value increases to 20dB to achieve BER of 10^{-3}. There is always trade-off between the bandwidth efficiency and channel estimation accuracy regards the number of pilot symbols. The number of pilot symbols will reduce the bandwidth efficiency but improves the channel estimation accuracy. Using ¼ the data the performance is improved but the bandwidth requirement for the pilots increase but using 1/16 the bandwidth efficiency increase but the performance of the estimator becomes worst.in this thesis work for extract the channel coefficients 1/8 of the data pilot is used.

5.2.6 Performance Analysis of MMSE and LS Channel Estimation in Different Multipath Channels

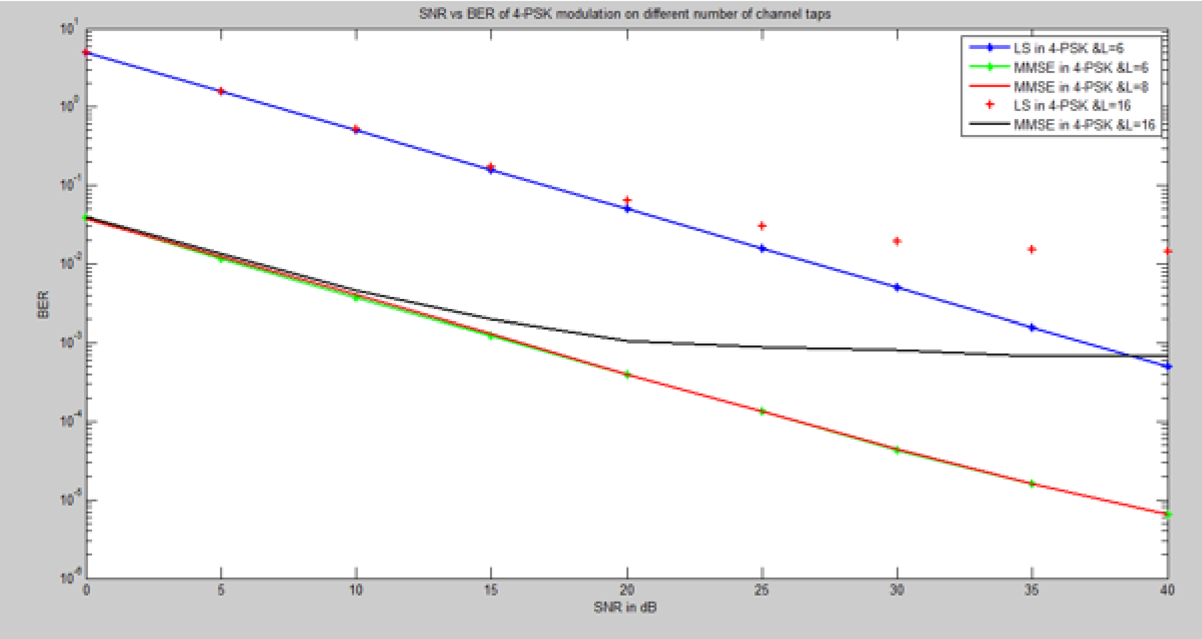


Figure 5.8 Performance of LS and MMSE estimators in different channel taps

Figure 5.8 indicates, as the number of channel taps or number of multipath increase the performance of the channel estimators becomes worst for both LS and MMSE estimator. The performance of MMSE estimator when the channel length is greater than the cyclic prefix becomes worst. In the figure above the black lines shows MMSE performance when the channel length is 16(CP> L ==>16> 8) this causes performance deviation in MMSE. But at the lower SNR the performance variation is very small.

5.2.7 Estimated channel weights

The aim of this thesis is to extract the channel fading coefficient for further channel equalization to combat ISI. The figure below shows the estimated channel coefficients using LS and MMSE estimators in QPSK modulation scheme.

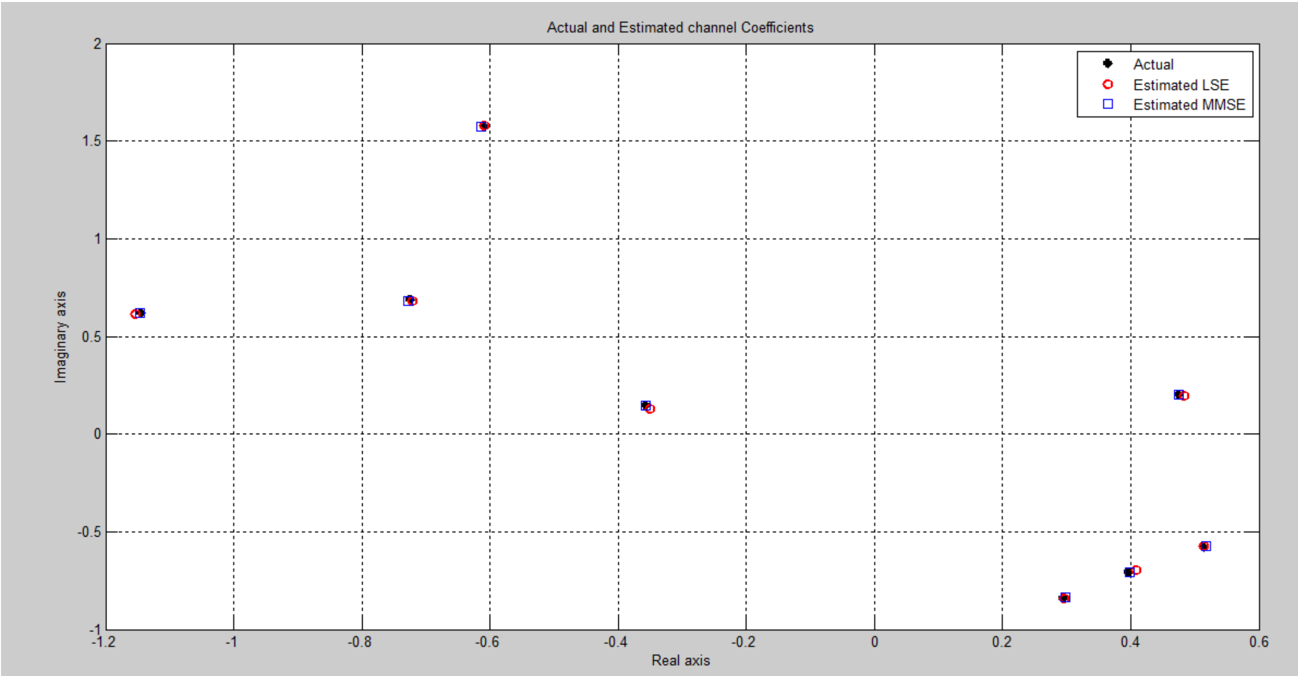


Figure 5.9 Actual and estimated channel coefficients

Figure 5.9 above shows the actual and estimated channel coefficients using MMSE and LS estimators. As we can see from the figure the estimated coefficients using MMSE estimator are closer to the actual coefficient than using LS estimator.

The extracted complex channel coefficient using LS and MMSE estimator and the average error is summarized in Table 5.2 below.

Table 5.2 estimated channel weights and errors in QPSK

	Actual channel	LS&MMSE Estimation algorithms results		Errors	
S. N	H(actual channel)	H_LS/estimated	H_MMSE/estimated	**H-H_LS**	**H-H_MMSE**
1	0.5149-0.5764i	0.5144-0.574i	0.5169-0.5744i	0.0005-0.0024i	-0.002-0.002i
2	-1.1449+0.6225i	-1.1537+0.6163i	-1.1469+0.6183i	0.0088+0.0062i	0.002+0.0042i
3	0.4752+0.199i	0.483+0.1966i	0.4744+0.2007i	-0.0078+0.0024i	0.0008-0.0017i
4	-0.3569+0.1474i	-0.3505+0.1318i	-0.3574+0.1462i	-0.0064+0.0156i	0.0005+0.0012i
5	-0.609+1.5739i	-0.6093+1.5767i	-0.6145+1.5715i	0.0003-0.0028i	0.0055+0.0024i
6	0.2952-0.8386i	0.2968-0.8387i	0.2981-0.8374i	-0.0016+0.0001i	-0.0029-0.0012i
7	-0.725+0.6844i	-0.7206+0.6786i	-0.7272+0.6817i	-0.0044+0.0058i	0.0022+0.0027i
8	0.3968-0.707i	0.4078-0.6973i	0.3992-0.7055i	-0.011-0.0097i	-0.0024-0.0015i
	Summation Errors			-0.0216+0.0152i	0.0037+0.0041i
	Average error			-0.0027+0.0019i	0.0004625+0.0005125i
	Error in linear number			0.0033	0.00069

From Table 5.2 above using LS algorithm the average error is about 3.3×10^{-3} and for MMSE algorithm the average error is about 6.9×10^{-4}.

5.2.8 Constellation Plot in Flat Fading channel

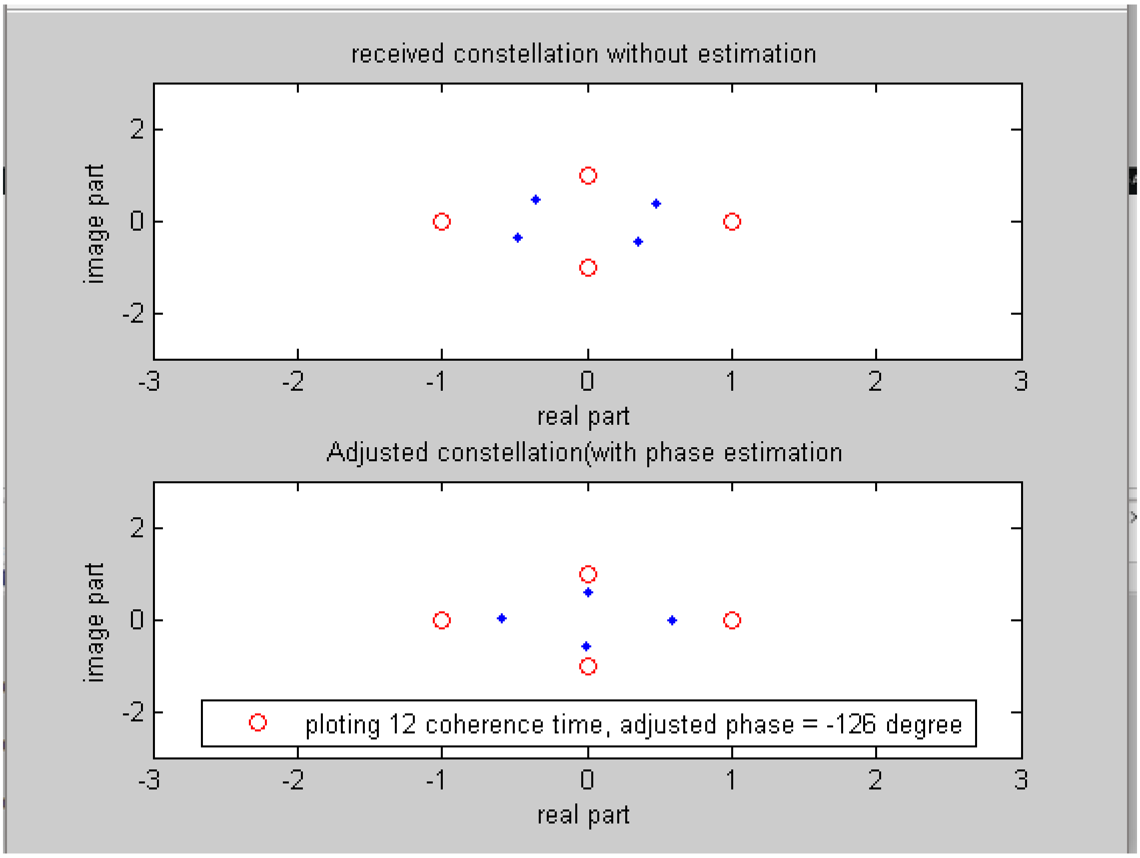


Figure 5.10 received and estimated constellation plot

In flat fading channel since there is no ISI the phase shifted can be adjusted using channel estimation. The Figure 5.10 above shows the received constellation is adjusted by 126 degree using channel estimator.

5.2.9 Received and Adjusted Image Comparison in Flat fading channel

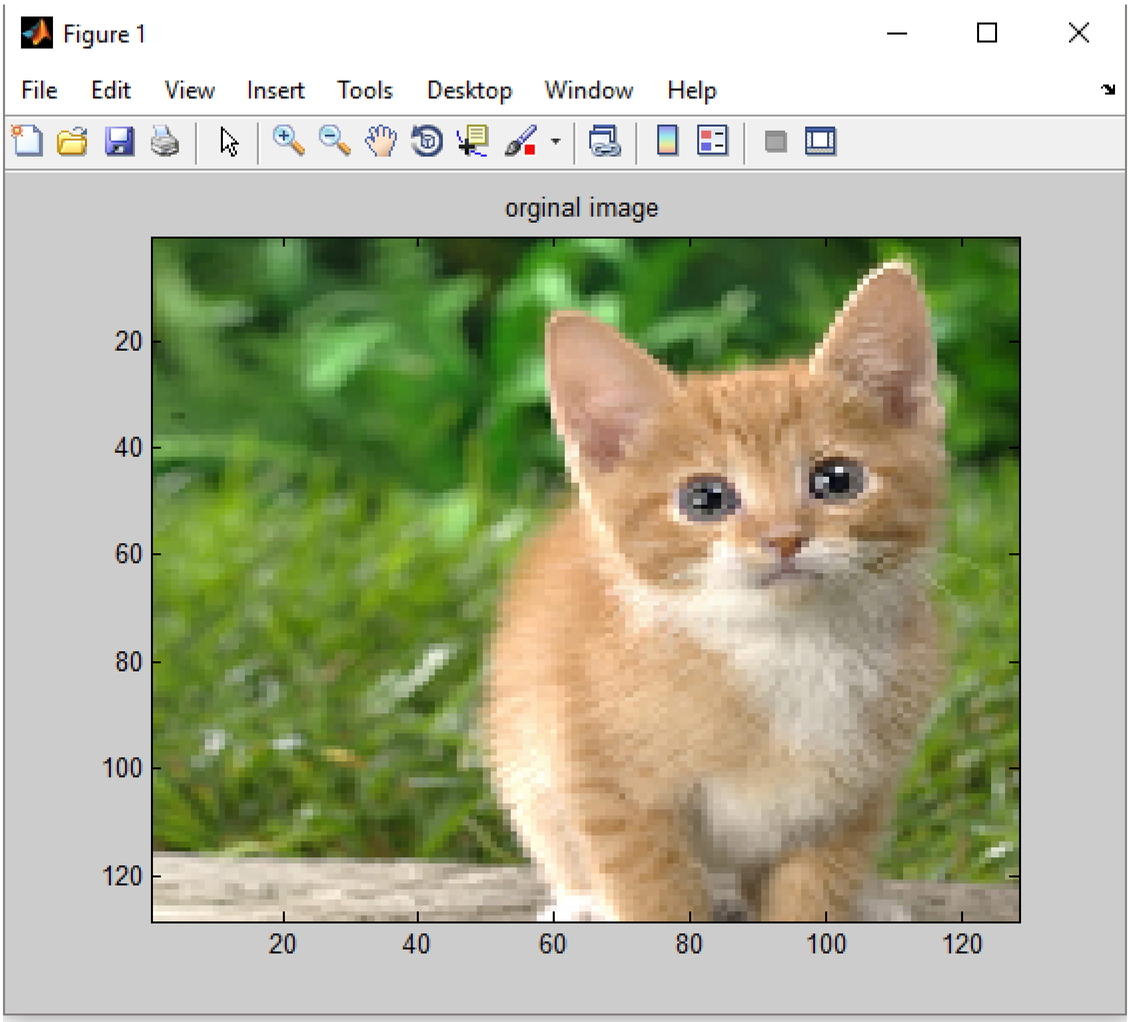
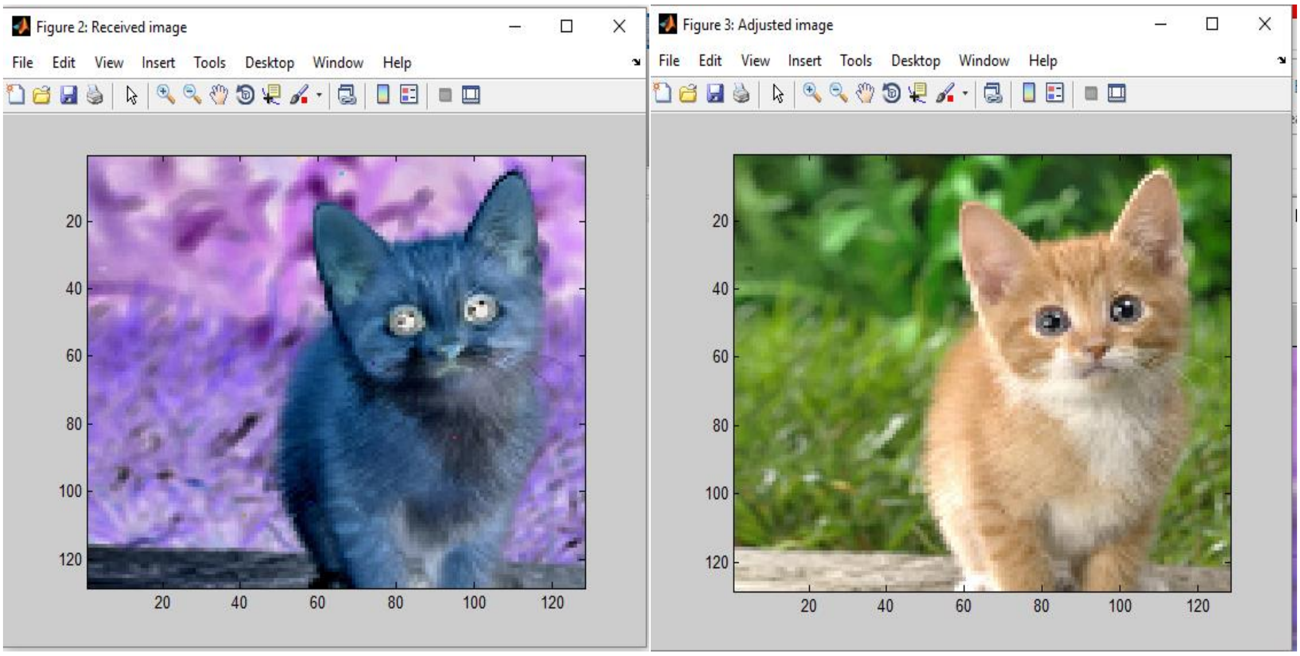


Figure 5.11 transmitted original image



a) Received Image b) Adjusted Image

Figure 5.12 received and adjusted image quality comparison

Figure 5.12 (a and b) shows the images received before and after estimation respectively. The received image without estimation is very erroneous and difficult to identify. But after channel estimation (phase adjustment) the adjusted image is almost the same as the original transmitted image. For this simulation QPSK modulation technique is used to test the phase shift recovery using LMS algorithm.

Chapter Six

Conclusions and Recommendations

6.1 Conclusions

In this thesis, a review of pilot-based channel estimation techniques for broadcasting system is given. Firstly, the general structure of channel estimation techniques and pilot-based channel estimation algorithms are reviewed. Then the performance of LS and the MMSE estimators, computational complexity are compared. Based on the performance analysis the MMSE estimator is recognized as better than LS estimator, but the MMSE estimator suffers from high computational complexity.

To extract the channel coefficients estimation is performed first at the pilot symbol positions using LS and MMSE estimator followed by channel interpolator to get the channel coefficients at the data symbols. For this purpose different types of interpolation techniques have been reviewed and their BER vs SNR performance is investigated. The simulation results show that the performance of the interpolator becomes better as the order of polynomial for interpolation increases. Based on the performance and complexity cubic spline interpolation is used in this thesis.

Iterative pilot-based estimators LMS and RLS have high speed of convergence but appropriate for joint estimation and equalization. So, the simulation is made only using LS and MMSE channel estimation algorithms to extract the channel coefficients. As it can be seen from Table 5.2 above using LS algorithm the MSE is about 3.3×10^{-3} and for MMSE algorithm the MSE is about 6.9×10^{-4} in QPSK modulation at 15dB.

6.2 Recommendations

There are a few open areas recommended for future works.

In this thesis the analysis is on pilot-based channel estimation techniques which is bandwidth inefficient. Semi-blind channel estimation techniques can improve the band width efficiency since it requires less pilot length than non-blind techniques and have better performance than totally blind channel estimation techniques. Therefore, performance of semi-blind channel estimator could be one area of future study.

In this analysis, all channels were assumed to be slow-fading and time selectivity is not considered. In a case of fast fading channels, channel fading coefficients vary within a symbol and the slow fading assumption does not hold. Therefore, performance of channel estimator in fast fading channel environments could be one area of future study.

In this thesis all the pilots are assumed as orthogonally arranged which is appropriate for one dimensional channel estimators i.e. frequency domain or time domain only. But two dimensional (doubly selective) estimations techniques that can be implemented using basis expansion model (BEM) and are more efficient but complexity increase.

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